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Power Rectification with Silicon Diodes

M. Dayal
Mullard Semiconductor Measurement and Application Laboratory

A descriptive survey is made of single-phase and three-phase rectifier circuits, ranging from a simple half-wave circuit to the double-star circuit with interphase reactor. The pros and cons of choke and capacitor input filters are discussed.

The potentialities of the possible circuit configurations, when equipped with silicon diode rectifiers, are examined in detail, and their usefulness for particular design applications is indicated. The power levels envisaged range from low levels to 1MW and more. Guidance is given on series and parallel operation and on protection against transients.

Design calculations and practical design procedures are included, with practical examples and performance figures; and there is a comprehensive reference table of rectifier circuit characteristics.

The work is limited to 'low frequency' operation (≤ 400c/s). A later article will discuss operation at higher frequencies, such as are required in high-frequency inverter and converter applications.

INTRODUCTION

The theory of rectification by means of vacuum tubes and mercury arc rectifiers has been established for many years. It is the object of the present article to show that the theory equally applies to silicon diode rectifiers, and that it is perfectly feasible to design high-power equipment using silicon rectifiers for powers up to a megawatt and more. Single-phase circuits are discussed in the first part (page 231), and three-phase circuits in the second (page 251). Parallel and series operation, and rectifier protection, are discussed in the final section (page 259).

Throughout the article it has been assumed that the rectifiers are operated at low frequencies; that is, at frequencies up to 400c/s, which is the frequency limit normally quoted for the published ratings.

In certain applications it is necessary to operate rectifiers at high frequency, as in inverter and converter circuits. Here the average current that may be passed through the rectifier must be derated from the low-frequency rating because of the heating effect caused by the minority carrier current flowing during the recovery period of the rectifier. High-frequency operation will be discussed in a later article.

A further assumption in the present article is that the rectifiers are ideal in the reverse direction. Leakage current has, however, been considered in the discussion of series operation.

A comprehensive table of rectifier circuit characteristics will be found on the centre pages of this issue.

SILICON RECTIFIERS

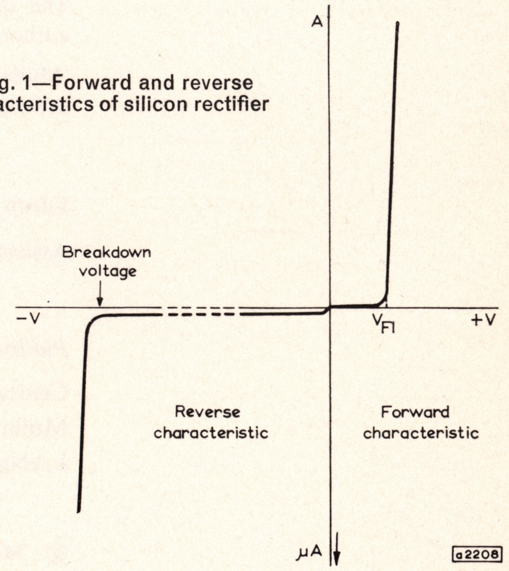
Apart from the more obvious advantages of size and weight, silicon diode rectifiers have proved to be efficient and reliable, and they require very little maintenance. However, they have limited overload current capacity and are sensitive to overvoltage peaks even of short duration. It is therefore necessary to choose rectifiers carefully for any particular application, bearing in mind the operational as well as transient surges that are likely to occur.

Characteristics

The voltage-current characteristic of a silicon rectifier is shown in Fig. 1. The forward characteristic shows that below a certain forward voltage V_{FI} only a very small current flows through the rectifier. This is because the current increases initially (to a first approximation) exponentially with the forward voltage. Beyond a certain voltage (normally between 0·5 and 1·0V) the characteristic becomes nearly linear, the current being almost solely limited by the differential resistance of this part of the characteristic. The differential resistance is of the order of hundredths of an ohm, and is sometimes even less.

In reverse, the rectifier presents a very high impedance, and therefore only a small current can flow in this direction (microamps or milliamps). The reverse characteristic is also shown in Fig. 1, to an expanded scale. The reverse current increases slightly with increase in reverse voltage. However, at a certain reverse voltage termed the breakdown voltage, the reverse current begins to increase rapidly; and any further increase in reverse voltage, however small, will lead to large reverse currents and the possible destruction of the device.

Fig. 1—Forward and reverse characteristics of silicon rectifier



The V_{FI} component of the forward voltage drop across the rectifier decreases slightly with increasing temperature, and the ohmic component increases. The reverse current and the breakdown voltage are temperature-sensitive, both the current and the voltage increasing with increase in temperature. However, for thermal stability reasons the rectifier reverse voltage rating is generally reduced at high temperatures. The breakdown voltage varies from one type of rectifier to another, and its value may be anything from hundreds of volts to a few kilovolts.

Voltage Ratings

The maximum non-repetitive peak reverse voltage rating of the rectifier is the voltage that must not be exceeded by any supply transient, and this is less than the breakdown voltage. The maximum repetitive peak reverse voltage rating must never be exceeded by any recurrent voltages in the circuit, and their duration must not exceed the limit given in the data sheet.

In addition, there is a specified crest reverse working voltage V_{RW} which was previously referred to as the recurrent peak inverse voltage (PIV). This is the crest value of the idealised sinewave applied to the rectifier.

The occurrence of transients on the supply mains (Ref. 1) must be taken into account by the user, since voltage peaks in excess of the maximum ratings of a semiconductor rectifier can destroy the device.

Thermal Considerations

The junction area of the silicon rectifier is small, therefore it operates at a high current density, and the thermal capacity of the device is small. In order to use rectifiers at high powers, it is necessary to apply some form of cooling system to conduct the heat away from the junction, and to ensure that the maximum mounting base temperature rating is not exceeded. Normally, this is achieved by bolting down one terminal of the device to a convection-cooled heatsink; but sometimes forced cooling is used for large equipments. In either case the cooling system must be designed to allow operation at the maximum required power and the desired ambient operating temperature without exceeding the maximum mounting base temperature rating.

The design of heatsinks is discussed in Ref. 2.

Single-phase Rectifier Circuits

RECTIFIER CIRCUITS WITH RESISTIVE LOAD

The commonly used single-phase rectifier circuits, and the output voltage waveforms for these circuits when used with a resistive load, are shown in Figs. 2 and 3 respectively. It should be noted that the diode symbol indicates *conventional* current flow, from anode to cathode.

The half-wave rectifier circuit conducts during the positive half cycle and blocks during the negative half cycle of the applied alternating voltage.

In the full-wave centre-tap circuit, the rectifiers are mounted so that rectifier A conducts when point (x) goes positive, and rectifier B conducts when point (y) goes positive.

In the full-wave bridge circuit, rectifiers 1 and 2 conduct during the positive half cycle, and rectifiers 4 and 3 during the negative half cycle.

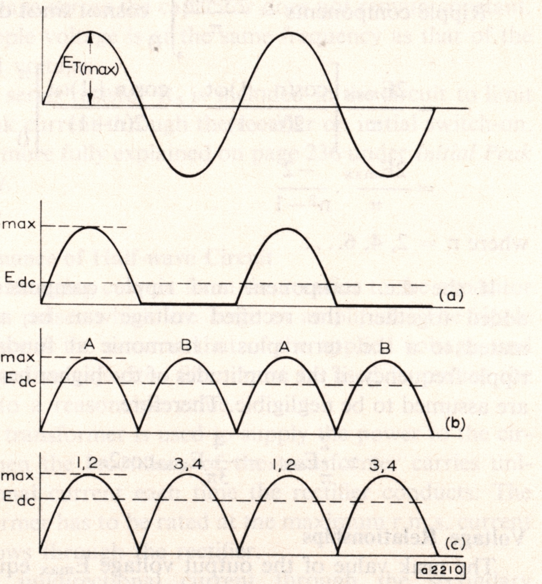
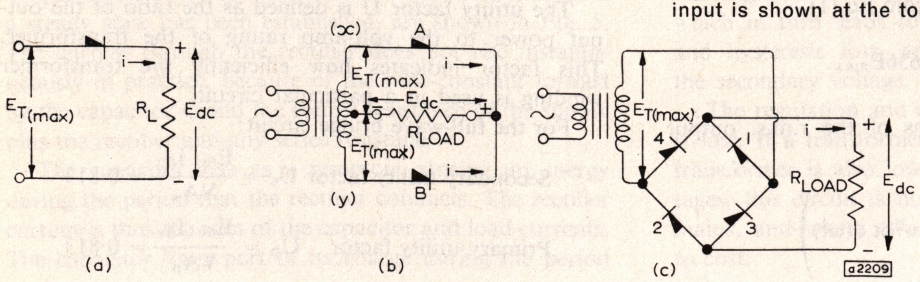


Fig. 3—Waveforms for single-phase circuits. The sinewave input is shown at the top, and the output waveforms below
(a) half-wave
(b) full-wave centre-tap
(c) full-wave bridge

Fig. 2—Single-phase rectifier circuits
(a) half-wave
(b) full-wave centre-tap
(c) full-wave bridge



The current through the load in each of the three circuits is unidirectional.

If it is assumed that the rectifiers and transformer used are ideal, the performance of any of these circuits can be calculated. The full-wave bridge circuit, for a resistive load, is calculated below as an example. The performance of other single-phase circuits can be derived in a similar way. The values obtained for each of the above circuits are given in Table 1 (page 246).

FULL-WAVE BRIDGE CIRCUIT PERFORMANCE WITH RESISTIVE LOAD

The sinusoidal rectified voltage from the bridge rectifier circuit may be expressed as a series containing a d.c. component and harmonic components.

$$e = \frac{2}{\pi}E_{max} - \frac{4}{3\pi}E_{max}\cos 2\omega t - \frac{4}{15\pi}E_{max}\cos 4\omega t - \frac{4}{35\pi}E_{max}\cos 6\omega t \ldots$$

This series for the rectified voltage is derived as follows:

The d.c. component is given by

$$E_{dc} = \frac{1}{2\pi}2\int_0^{\pi} E_{max}\sin\omega t \, d(\omega t) = \frac{2}{\pi}E_{max}.$$

The value of the ripple components, of frequency $n\omega/2\pi$, can be derived by means of Fourier series.

$$\text{Ripple components} = \frac{2E_{max}}{\pi}\int_0^{\pi} \cos n\omega t \sin\omega t \, d(\omega t)$$

$$= \frac{2E_{max}}{\pi}\left[\frac{\cos(n-1)\omega t}{2(n-1)} - \frac{\cos(n+1)\omega t}{2(n+1)}\right]_0^{\pi}$$

$$= \frac{2E_{max}}{\pi}\cdot\frac{-2}{n^2-1}$$

where $n = 2, 4, 6\ldots.$

If the d.c. component and ripple components are added together, the rectified voltage can be approximated to a d.c. term plus a harmonic at fundamental ripple frequency, if the amplitudes of the higher harmonics are assumed to be negligible. Therefore

$$e \simeq \frac{2}{\pi}E_{max} - \frac{4}{3\pi}E_{max}\cos 2\omega t. \qquad \ldots(1)$$

Voltage Relationships

The peak value of the output voltage E_{max} equals the peak value of the input voltage $E_{T(max)}$. The output voltage E_{dc} in terms of E_{max} is, from Eq (1).

$$E_{dc} = \frac{2}{\pi}E_{max} = 0{\cdot}636E_{max}.$$

The output voltage E_{dc} in terms of the r.m.s. output voltage E_{rms} is derived from

$$E_{rms} = \sqrt{\left(\frac{1}{2\pi}\int_0^{2\pi} E^2_{max}\sin^2\omega t \, d(\omega t)\right)}$$

$$= E_{max}\sqrt{\left(\frac{1}{2\pi}\int_0^{2\pi}\left(\frac{1}{2} - \frac{\cos 2\omega t}{2}\right)d(\omega t)\right)}$$

$$= \frac{E_{max}}{\sqrt{2}},$$

so that

$$E_{dc} = 0{\cdot}636\sqrt{2}E_{rms} = 0{\cdot}9E_{rms}.$$

Current Relationships

The output current waveform is the same as the output voltage waveform for resistive load.

The output current I_{dc} in terms of I_{pk} is given by

$$I_{dc} = \frac{2}{\pi}\frac{E_{max}}{R_{LOAD}} = 0{\cdot}636I_{pk}.$$

The output current I_{dc} in terms of $I_{rms(total)}$ is given by

$$I_{dc} = 0{\cdot}9\frac{E_{rms}}{R_{LOAD}} = 0{\cdot}9I_{rms(total)}.$$

The total direct current is supplied by two pairs of rectifiers, so that the average current per rectifier leg is

$$I_0 = \tfrac{1}{2}I_{dc}.$$

The total r.m.s. current is supplied by two pairs of rectifiers, so that the r.m.s. current per rectifier leg is

$$I_{rms} = \frac{I_{rms(total)}}{\sqrt{2}} = \frac{I_{dc}}{0{\cdot}9\sqrt{2}} = 0{\cdot}785I_{dc}.$$

The peak current per rectifier leg is

$$I_{pk} = \frac{E_{max}}{R_{LOAD}} = \frac{I_{dc}}{0{\cdot}636} = 1{\cdot}57I_{dc}.$$

Crest Working Voltage

The total transformer secondary r.m.s. voltage per leg is given by

$$E_{T(rms)} = \frac{E_{dc}}{0{\cdot}9} = 1{\cdot}11E_{dc},$$

and the crest working voltage is

$$V_{RW} = \sqrt{2}E_{T(rms)} = 1{\cdot}57E_{dc}.$$

In terms of E_{rms},

$$V_{RW} = \sqrt{2}E_{rms} = 1{\cdot}414E_{rms} \; (= 1{\cdot}414E_{T(rms)}).$$

Transformer Rating

The transformer secondary r.m.s. current is given by

$$I_{T(rms)} = \sqrt{2}I_{rms} = 1{\cdot}11I_{dc}.$$

The VA product of the secondary winding is

$$VA_s = E_{T(rms)}.I_{T(rms)} = 1{\cdot}23E_{dc}.I_{dc}.$$

If the primary to secondary transformer ratio is N_p/N_s, then the VA product of the primary winding is

$$VA_p = E_{T(rms)}\frac{N_p}{N_s}I_{T(rms)}\frac{N_s}{N_p}$$

$$= E_{T(rms)}.I_{T(rms)} = 1{\cdot}23E_{dc}.I_{dc}.$$

The utility factor U is defined as the ratio of the output power to the volt-amp rating of the transformer. This factor indicates how efficiently the transformer winding is used in a particular circuit.

For the full-wave bridge circuit:

$$\text{Secondary utility factor } U_s = \frac{E_{dc}.I_{dc}}{VA_s} = 0{\cdot}813$$

$$\text{Primary utility factor } \quad U_p = \frac{E_{dc}.I_{dc}}{VA_p} = 0{\cdot}813.$$

Percentage Ripple

If it is assumed that the amplitudes of the higher harmonics are small compared to that of the harmonic at fundamental frequency f_r, then

$$V_R\% = \frac{\text{Fundamental r.m.s ripple voltage}}{E_{dc}} \times 100.$$

From Eq (1), the r.m.s. harmonic component at fundamental frequency (which is twice the supply frequency for this circuit) is

$$\frac{4}{3\pi}\cdot\frac{E_{max}}{\sqrt{2}}$$

therefore

$$V_R\% = \frac{\frac{4}{3\pi}\cdot\frac{E_{max}}{\sqrt{2}}}{\frac{2}{\pi}E_{max}} \times 100 = 47{\cdot}2.$$

SINGLE-PHASE CIRCUITS WITH CAPACITOR INPUT FILTER

Single-phase half-wave, full-wave, and voltage-doubler circuits are discussed in this section.

Half-wave Circuit

The half-wave circuit, Fig. 4, is the simplest rectification circuit giving continuous load current. In the absence of the capacitor C the rectifier will deliver power to the load R_{LOAD} during the positive half cycle, and will block during the negative half cycle. This leads to discontinuous voltage and current in the load.

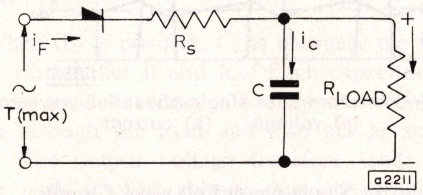


Fig. 4—Single-phase half-wave circuit

With the capacitor C in circuit, the capacitor charges to the crest value of the applied voltage on the first positive half cycle. When the applied voltage falls below the crest value, the capacitor voltage is higher than the applied voltage, and thus the rectifier is reverse biased. The capacitor now discharges into the load until such time as the applied voltage exceeds the capacitor voltage again. The rectifier is then forward biased and charges the capacitor to the crest applied voltage again. The rectifier then ceases to conduct, as previously explained, and the cycle is repeated.

The idealised current waveforms for this circuit, after a steady state has been established, are shown in Fig. 5. The current through the rectifier does not rise instantaneously in practice, because of the time-constant formed by the capacitor C and the source resistance of the supply plus the rectifier and any series resistance.

The capacitor acts as a reservoir, storing up energy during the period that the rectifier conducts. The rectifier current is thus the sum of the capacitor and load currents. The capacitor loses part of its charge during the period that the rectifier is non-conducting, because it discharges into the load during this period. The load current is then equal to the capacitor current i_l. Because of this action, the voltage across the capacitor does not remain constant. The ripple voltage is at the same frequency as that of the applied voltage.

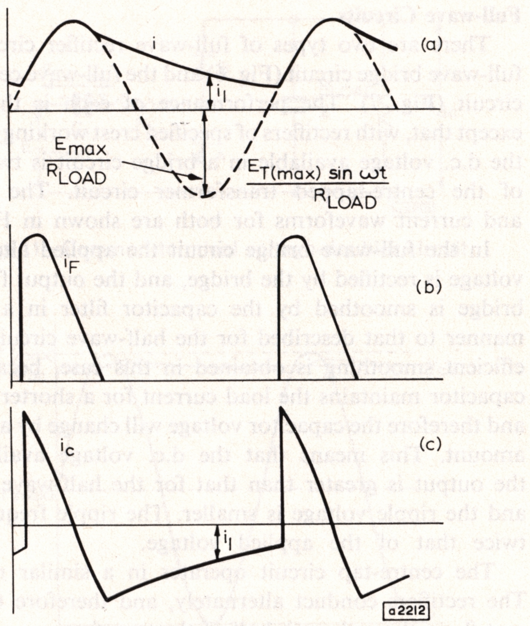


Fig. 5—Waveforms for single-phase half-wave circuit, after establishment of steady state
(a) load current
(b) rectifier current
(c) capacitor current

The series resistor R_s is included in the circuit to limit the peak current through the rectifier on initial switch-on. This is more fully explained on page 236 under *Initial Peak Current.*

Performance of Half-wave Circuit

The charging current from the rectifier to the capacitor flows in pulses which are large in amplitude. The ripple frequency is the same as that of the applied voltage, and an expensive capacitor filter must be used to reduce the ripple to a reasonable value.

If a transformer is used to supply the power to the circuit, then the secondary of the transformer carries unidirectional current each time the rectifier conducts. The transformer has to be rated at the maximum r.m.s. current that flows through the rectifier.

The unidirectional current through the secondary winding of the transformer can lead to core saturation, which in turn leads to increases in magnetising current and hysteresis loss, and introduction of harmonics in the secondary voltage.

The regulation and conversion efficiency of the circuit is low. If a transformer is used, the utility factor of the transformer is also low. Because of the above disadvantages, this circuit is normally only used direct from the mains, and where efficiency is of secondary importance to cost.

Full-wave Circuits

There are two types of full-wave rectifier circuit: the full-wave bridge circuit (Fig. 6) and the full-wave centre-tap circuit (Fig. 7). The performance of each is the same, except that, with rectifiers of specified crest working voltage, the d.c. voltage available in a bridge circuit is twice that of the centre-tapped transformer circuit. The voltage and current waveforms for both are shown in Fig. 8.

In the full-wave bridge circuit the applied alternating voltage is rectified by the bridge, and the output from the bridge is smoothed by the capacitor filter in a similar manner to that described for the half-wave circuit. More efficient smoothing is obtained in this case, because the capacitor maintains the load current for a shorter period, and therefore the capacitor voltage will change by a smaller amount. This means that the d.c. voltage available at the output is greater than that for the half-wave circuit, and the ripple voltage is smaller. The ripple frequency is twice that of the applied voltage.

The centre-tap circuit operates in a similar manner. The rectifiers conduct alternately, and therefore the current flows through each half of the transformer secondary alternately. The rectifiers must withstand a crest working voltage which is equal to the peak value of the applied voltage across both halves of the transformer secondary.

Comparison of Single-phase Full-wave Circuits

The two full-wave circuits are compared in Table 2.

TABLE 2

Comparison of Single-phase Full-wave Circuits

	Bridge (Fig. 6)	Centre-tap (Fig. 7)
No. of rectifiers	4	2
Ripple frequency f_r	$f_r = 2f$	$f_r = 2f$
Ripple amplitude	Small compared to half-wave circuit	Small compared to half-wave circuit
Smoothing	Relatively easy	Relatively easy
Crest working voltage	$E_{T(max)}$	$2E_{T(max)}$
Conversion efficiency	Relatively high, but slightly lower than for centre-tap circuit because of voltage drop across additional rectifier	High
Transformer	Low transformer secondary volt-amp rating	High transformer secondary volt-amp rating

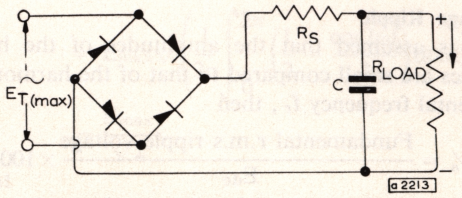


Fig. 6—Single-phase full-wave bridge circuit

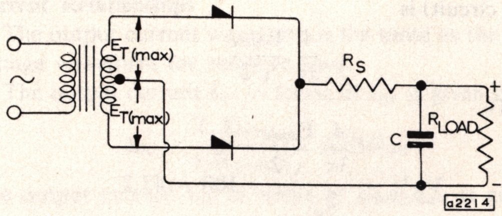


Fig. 7—Single-phase full-wave centre-tap circuit

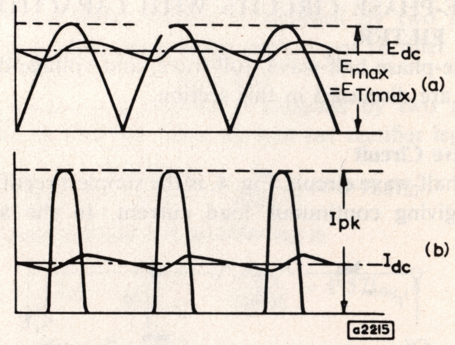


Fig. 8—Waveforms for single-phase full-wave circuits
(a) voltage (b) current

Applications of Single-phase Full-wave Circuits

The principal drawback of the centre-tap circuit is the cost of the transformer. The circuit can never be used without a transformer, whereas in certain circumstances the full-wave bridge circuit may be directly operated from the mains, if the rectifiers used are rated to withstand the crest working voltage. On the other hand, it is easy to obtain a three-wire d.c. supply from a single transformer with the centre-tap circuit.

The bridge circuit is the more widely used of the full-wave circuits. It is generally used wherever the desired output voltage is approximately equal to the r.m.s. applied voltage. The centre-tap full-wave circuit is used for low-power and low-voltage applications, where low ripple is desired.

Voltage-Doubler Circuits

There are two types of voltage-doubler circuit: the symmetrical and common-terminal circuits.

Symmetrical Voltage-doubler

The symmetrical voltage-doubler (Fig. 9) is essentially a combination of two half-wave rectifier circuits, with smoothing filters connected in series, but supplied from the same power source. The output voltage waveform is shown in Fig. 10.

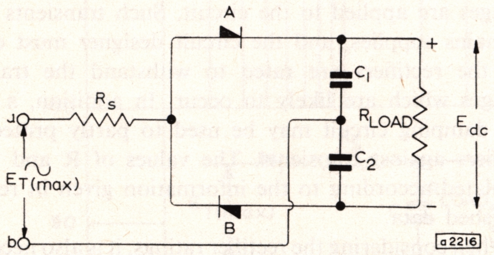


Fig. 9—Symmetrical voltage-doubler circuit

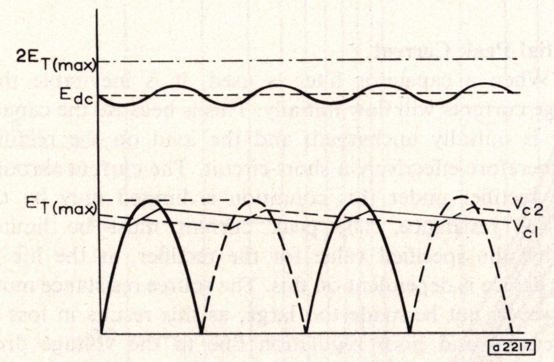


Fig. 10—Output voltage waveform for symmetrical voltage-doubler

When (a) is positive, current flows through R_s and rectifier A to charge C_1, whose other terminal is connected to (b). When (b) is positive, C_2 is charged, the return to (a) being via rectifier B and R_s. Each capacitor charges to the peak applied voltage. The capacitors continually discharge through the load and also act as smoothing elements. The output voltage therefore tends towards twice the peak applied voltage, but cannot achieve it unless the load R_{LOAD} is disconnected.

The rectifiers must be able to withstand twice the peak applied voltage in the reverse direction. The capacitors must be rated at the peak applied voltage. The ripple frequency of the circuit is twice that of the applied voltage.

Common-terminal Voltage-doubler

The common-terminal voltage-doubler is shown in Fig. 11, and the output voltage waveform in Fig. 12.

During the first negative half cycle of the applied voltage, C_1 charges to the peak voltage $E_{T(max)}$ through rectifier A. During the next positive half cycle, the voltage across C_1 is in series with the applied voltage and aids it to charge C_2 to $2E_{T(max)}$ through rectifier B. Capacitor C_1 loses part of its charge during this process, but charges up to $E_{T(max)}$ again during the next negative half cycle. The cycle is then repeated.

The voltage across C_2 does not remain constant at approximately $2E_{T(max)}$ because it discharges into the load R_{LOAD} when rectifier B is not supplying the load current.

The ripple frequency is the same as that of the applied voltage. Capacitor C_2 must be rated at twice the peak applied voltage, and the rectifiers must withstand twice the peak applied voltage.

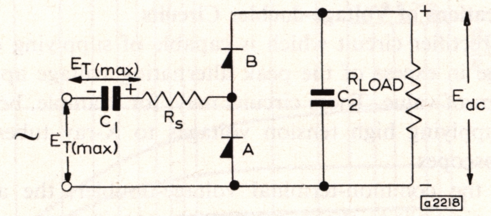


Fig. 11—Common-terminal voltage-doubler circuit

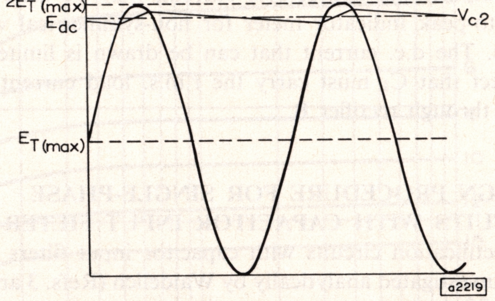


Fig. 12—Output voltage waveform for common-terminal voltage-doubler

Applying similar reasoning, it is relatively simple to construct a voltage tripler, a voltage quadrupler, or a circuit with an output voltage which is any other multiple of the peak applied voltage. The essential points to bear in mind are (a) the maximum crest working voltage that the rectifier must withstand, and (b) the rating of the capacitors.

Comparison of Voltage-Doubler Circuits

The two voltage-doubler circuits are compared in Table 3.

TABLE 3

Comparison of Single-phase Voltage-doubler Circuits

	Symmetrical	Common-terminal
Crest working voltage	$2E_{T(max)}$	$2E_{T(max)}$
Ripple frequency f_r	$f_r = 2f$	$f_r = f$
Capacitor rating	Rating of C_1 and C_2 must be equal to the peak applied voltage.	Rating of C_1 must be equal to the peak applied voltage, and that of C_2 twice the peak applied voltage. C_1 must be rated to carry the r.m.s. load current
Regulation	Poor, but better than for common-terminal doubler	Poor

Applications of Voltage-doubler Circuits

A rectifier circuit which is capable of supplying direct voltage in excess of the peak alternating voltage applied, is often of value. These circuits may, for example, be used for supplying high tension voltages to X-ray tubes and oscilloscopes.

In the common-terminal voltage-doubler, the action of the circuit is to charge the output capacitor to a voltage equal to the sum of the voltage across C_1 and the peak positive applied voltage. Therefore this circuit, together with some means of measuring the direct output voltage without drawing appreciable current, may be used as a peak-to-peak indicator meter for non-symmetrical waveforms. The d.c. current that can be drawn is limited by the fact that C_1 must carry the r.m.s. load current that flows through rectifier B.

DESIGN PROCEDURE FOR SINGLE-PHASE CIRCUITS WITH CAPACITOR INPUT FILTER

Rectification circuits with capacitor input filters have been investigated analytically by Waidelich (Refs. 3 and 4), Roberts (Ref. 5), and many others, and graphically by Schade (Ref. 6). Both methods of analysis are long and involved; but, fortunately, they lead to results that are simple to use. The above analyses were originally carried out on mercury vapour diodes and high-vacuum diodes, but they apply to silicon rectifiers with little or no modification.

The graphical analysis carried out by Schade has proved over many years to yield sufficiently accurate results; moreover, the design procedure is by far the simplest, and forms the basis of the procedure given below. The design charts have been adapted, by permission, from the original article by Schade.

The forward voltage drop across the silicon rectifier is small, and varies by a small amount with forward current. The increase in forward voltage may be ignored without any loss of accuracy because in most circuits this represents a very low percentage of the output voltage. Therefore, for the purpose of calculations, the forward voltage drop may be considered to be that voltage drop which occurs across the rectifier when the maximum required average current is flowing through the rectifier.

Factors to be Considered

When designing any rectifier circuit it is necessary to ensure that the published rectifier ratings are not exceeded. The four main characteristics to note in circuits with capacitor input filter are:

(i) Maximum crest working voltage rating of the rectifier
(ii) Initial switch-on peak current through the rectifier
(iii) Recurrent peak current through the rectifier
(iv) Ripple current through the capacitor.

Maximum Crest Working Voltage

The rectifier will withstand the crest working voltage for which it is rated, when an alternating voltage is applied. It will also withstand the peak transient voltage (in general higher than the crest working voltage) if transient voltages are applied to the circuit. Such transients occur on mains supplies, and the circuit designer must ensure that the rectifiers are rated to withstand the transient voltages which are likely to occur. In addition, a series R-C damping circuit may be used to partly protect the rectifiers against transients. The values of R and C are calculated according to the information given in rectifier published data.

When considering the rectifier ratings, it is also necessary to take into account the fluctuation in the alternating voltage, and the input waveform distortion due to harmonics.

Initial Peak Current

When a capacitor filter is used, it is inevitable that large currents will flow initially. This is because the capacitor is initially uncharged, and the load on the rectifier is therefore effectively a short-circuit. The current through the rectifier under this condition is limited only by the source resistance. The peak current must be limited below the specified value for the rectifier, as the life of the device is dependent on this. The source resistance must, however, not be made too large, as this results in loss of efficiency and poor regulation due to the voltage drop across it.

Recurrent Peak Current

The recurrent peak current flows through the rectifier each time it conducts. Its value is dependent on the value of the reservoir capacitor. Improvement in smoothing (increasing C) results in a reduction in the rectifier conduction angle, and an increase in the recurrent peak current. This recurrent peak current must be limited below the specified value if the life of the rectifier is not to be impaired.

Ripple Current

The capacitor used in the circuit must be rated to handle the ripple current that will flow through it. The total r.m.s. current flowing through the reservoir capacitor, $I_{c(rms)}$, can be calculated from the r.m.s. current flowing through each rectifier, I_{rms}, and the d.c. output current I_{dc}.

For single-phase half-wave and voltage-doubler circuits

$$I_{c(rms)} = \sqrt{(I_{rms}{}^2 - I_{dc}{}^2)}. \qquad \ldots(2)$$

In the full-wave rectifier circuits, half the total r.m.s. current flows through each rectifier, therefore

$$I_{c(rms)} = \sqrt{(2I_{rms}{}^2 - I_{dc}{}^2)}. \qquad \ldots(3)$$

From the above considerations, it is clear that the rectifier circuits using capacitor filters are limited in their current-handling capacity.

GRAPHICAL DESIGN OF CIRCUITS USING CAPACITOR INPUT FILTERS

The graphical solution of a capacitor filter rectifier circuit, as put forward by Schade, is presented in Figs. 13 to 18. The peak resistance $\hat{R}_s$ introduced by Schade to include the peak tube resistance is replaced by the source resistance R_s, which includes the transformer winding resistance, the rectifier resistance, and the series resistance added to limit the initial peak rectifier current.

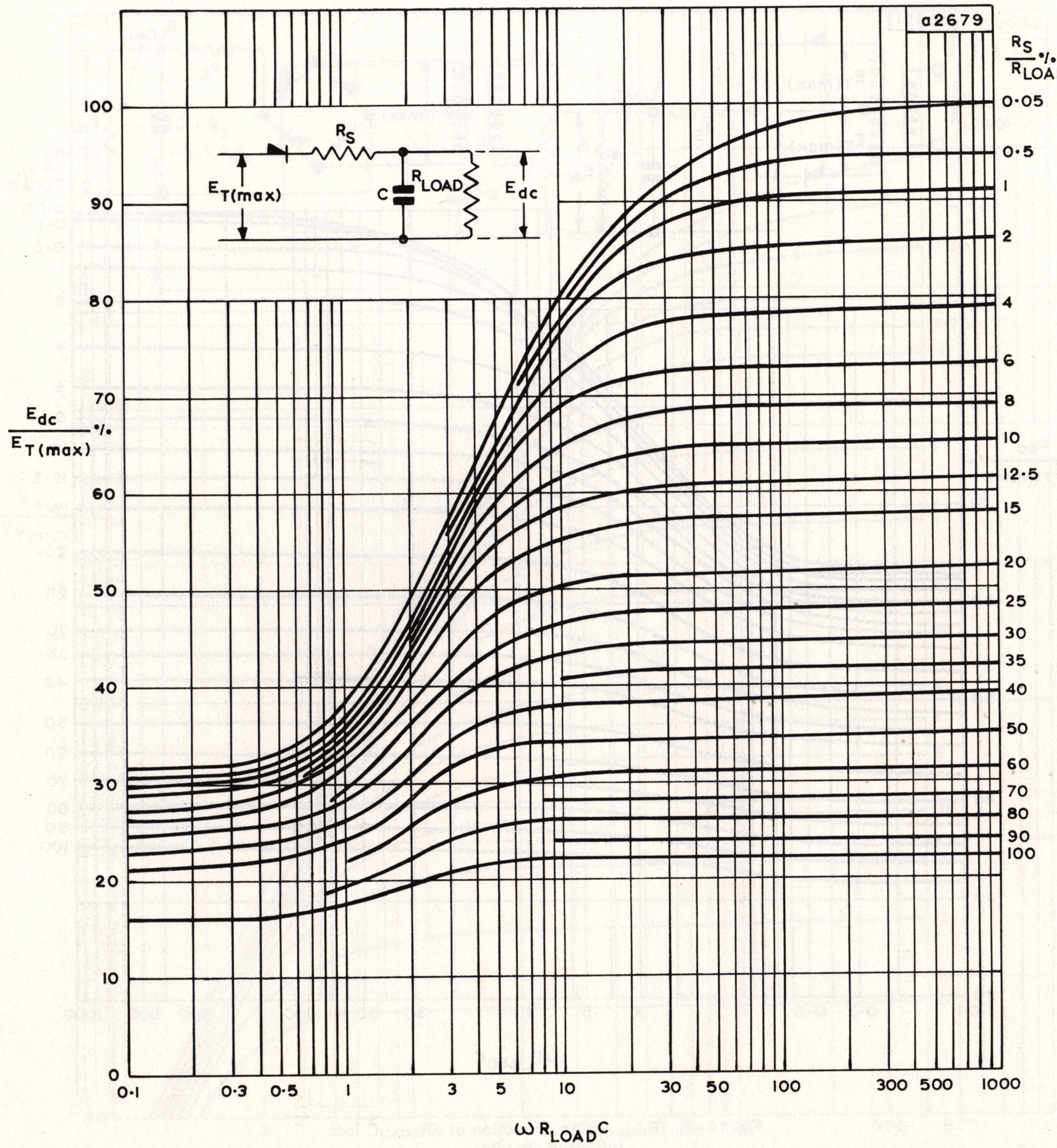


Fig. 13—$E_{dc}/E_{T(max)}$% as a function of $\omega R_{LOAD}C$ for half-wave circuits.
C in F, and R_{LOAD} in Ω $\omega = 2\pi f$

Figs. 13, 14, and 15 give the conversion ratio $E_{dc}/E_{T(max)}$ as a function of $\omega R_{LOAD}C$ for half-wave, full-wave, and voltage-doubler circuits respectively. The conversion ratio depends on the value of (R_s/R_{LOAD}%). For reliable operation the value of $\omega R_{LOAD}C$ should be selected to allow operation on the flat portion of the curves.

Fig. 16 gives information on the minimum value of $\omega R_{LOAD}C$ that must be used to reduce the percentage ripple to a desirable figure. Figs. 17 and 18 give, respectively, the ratio of r.m.s. rectifier current to average current per

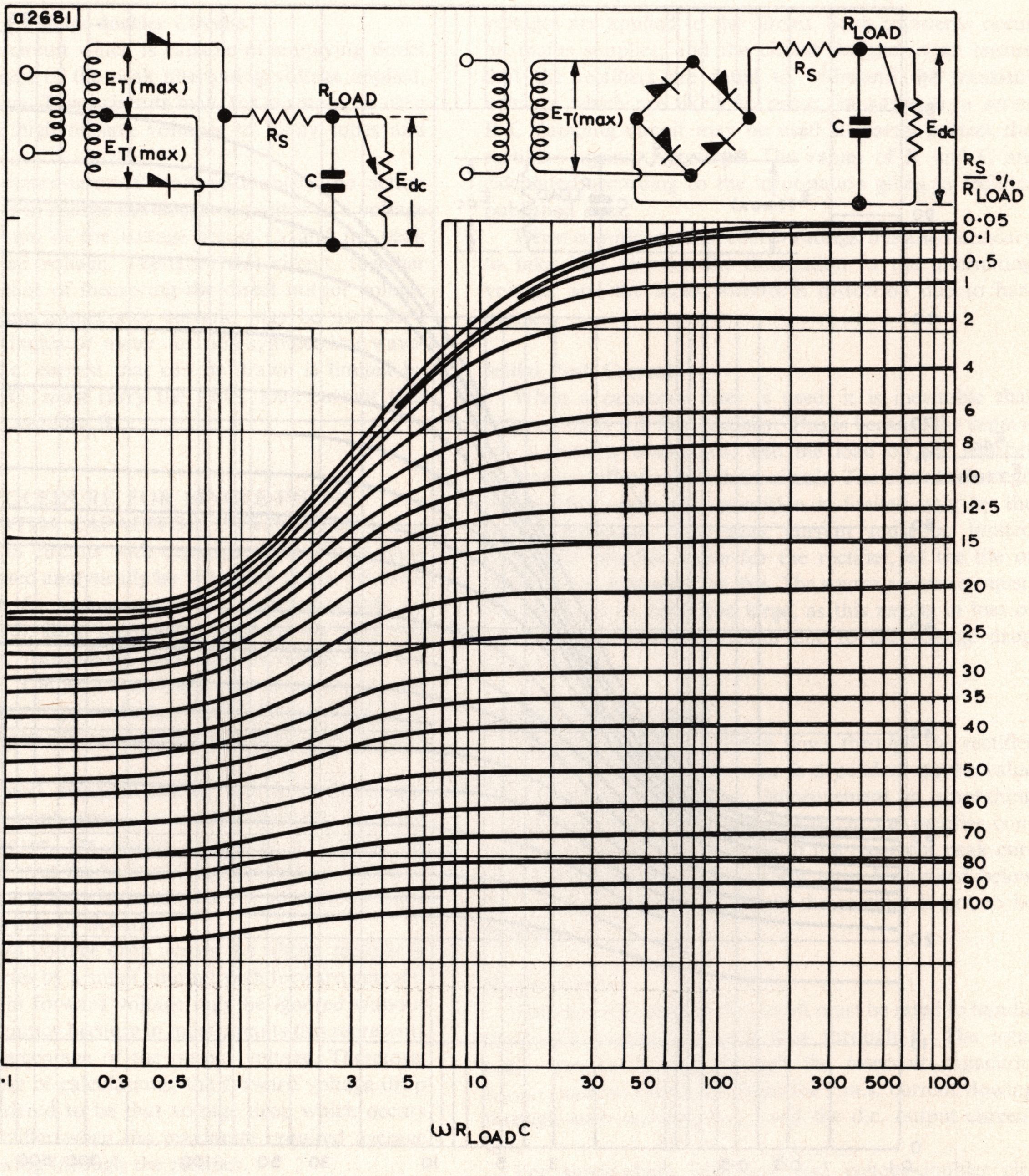


Fig. 14—$E_{dc}/E_{T(max)}$% as a function of $\omega R_{LOAD}C$ for full-wave circuits.
C in F, and R_{LOAD} in Ω $\omega = 2\pi f$

rectifier and the ratio of peak recurrent rectifier current to average current per rectifier, plotted as functions of $n\omega R_{LOAD}C$. These ratios are dependent on the value of R_s/nR_{LOAD}%.

The transformer leakage reactance has not been taken into account in the design procedure. However, it tends to reduce the peak rectifier current, and therefore assists in limiting the peak current.

Design Procedure

The following design procedure is recommended in the design of single-phase silicon rectifier circuits with capacitor input filter.

(1) Determine the value of R_{LOAD}.

(2) Assume a value of R_s (usually between 1 and 10% of R_{LOAD}).

(3) Calculate R_s/R_{LOAD}%.

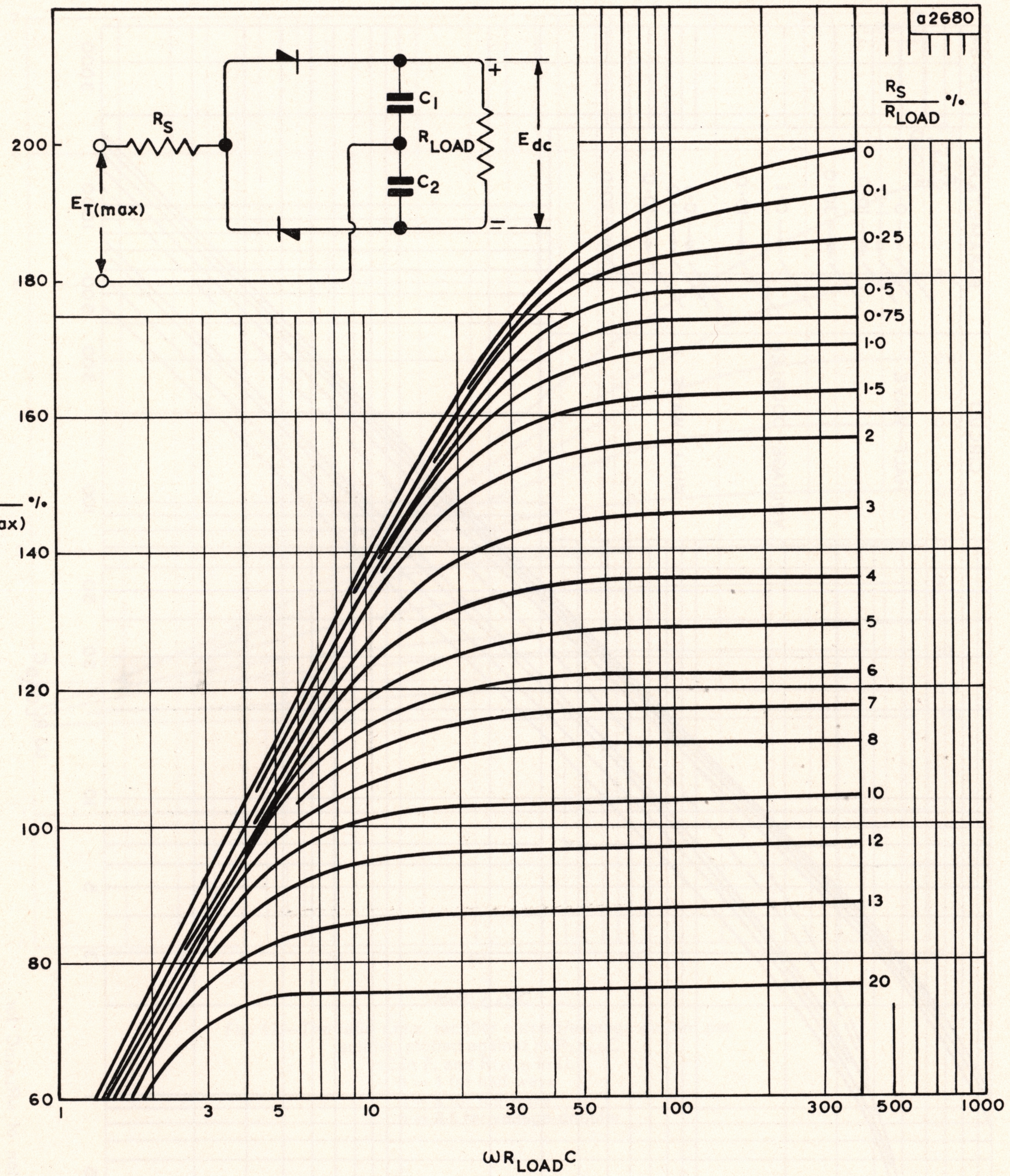


Fig. 15—$E_{dc}/E_{T(max)}$% as a function of $\omega R_{LOAD}C$ for voltage-doubler circuits.
C in F, and R_{LOAD} in Ω $\omega = 2\pi f$

(4) From the percentage ripple graph against $\omega R_{LOAD}C$ (Fig. 16), determine the value of $\omega R_{LOAD}C$ required to reduce the ripple to a desired value for R_s/R_{LOAD}% determined in (3). Calculate the value of C required.

(5) From the $E_{dc}/E_{T(max)}$% against $\omega R_{LOAD}C$ curves for the appropriate circuit (Fig. 13, 14, or 15) determine the conversion ratio for the value of $\omega R_{LOAD}C$ determined in (4) and R_s/R_{LOAD}% determined in (3).

(6) Determine the $E_{T(max)}$ and $E_{T(rms)}$ that must be applied to the circuit, using information derived in (5).

(7) Determine the crest working voltage that the rectifiers must withstand.

(8) Determine the r.m.s. current per rectifier from Fig. 17.

(9) Decide on the rectifiers to be used.

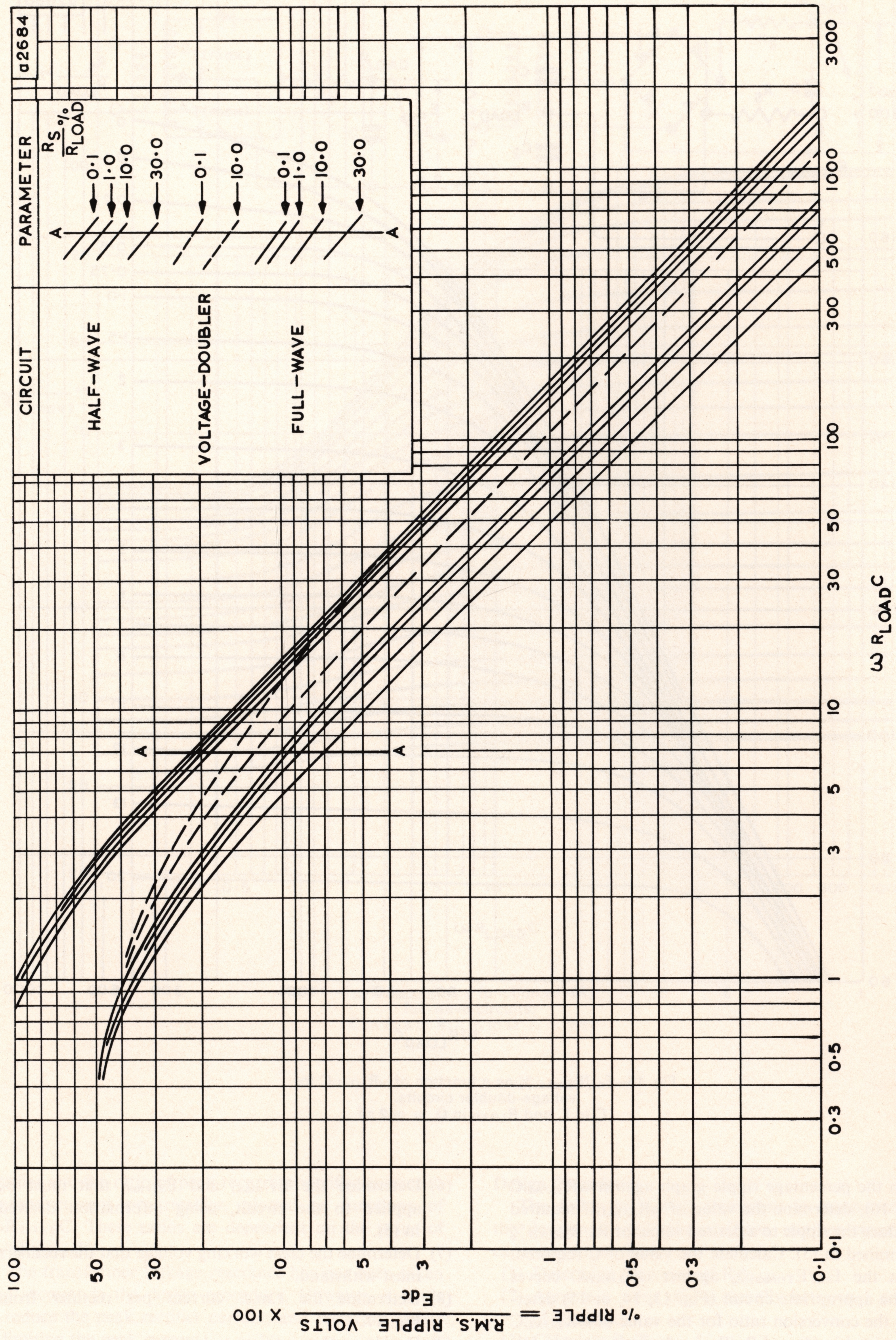


Fig. 16—Percentage ripple as a function of $\omega R_{LOAD}C$ for capacitor input filter.
C in F, and R_{LOAD} in Ω.
$\omega = 2\pi f$
f = line frequency

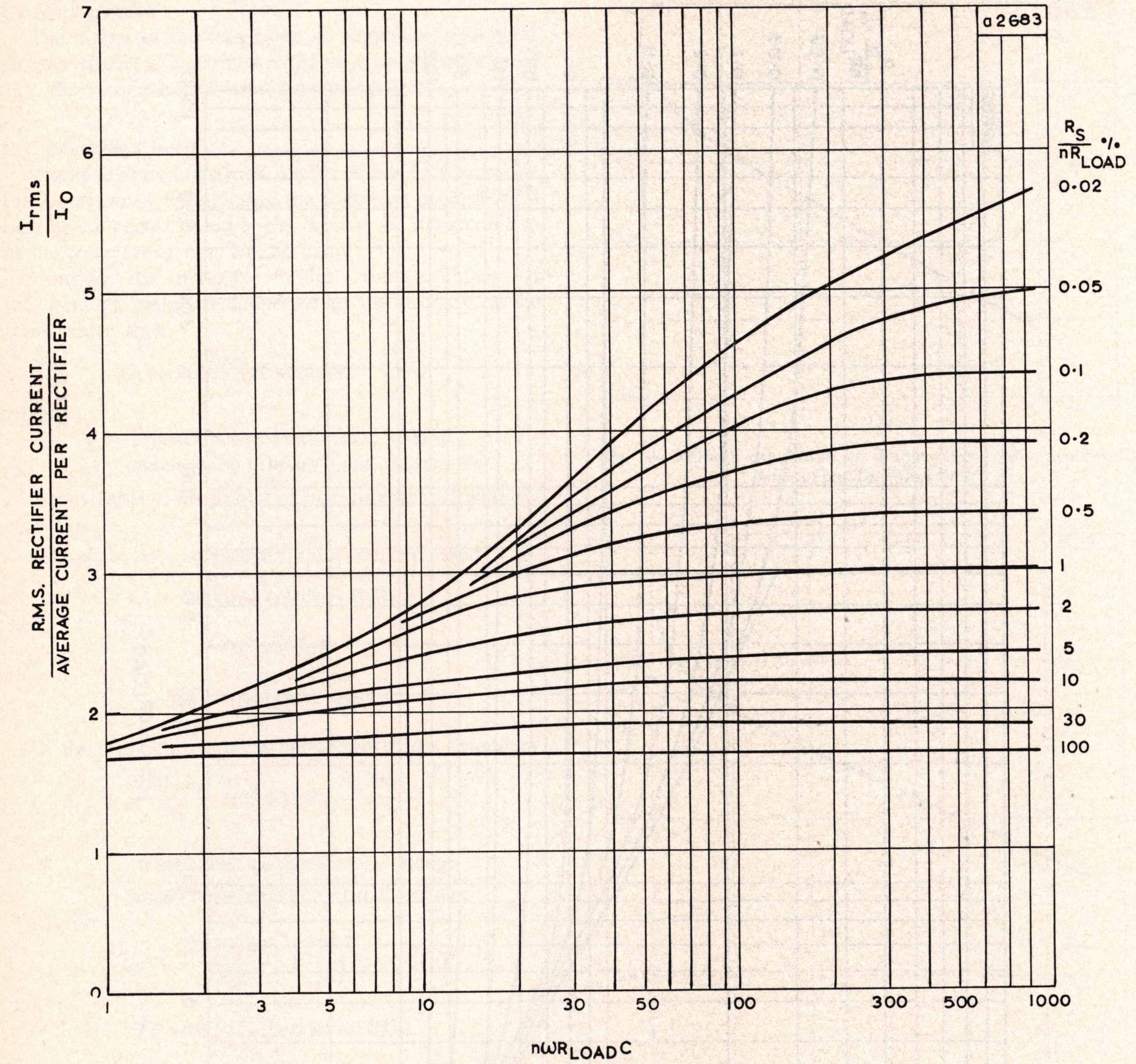


Fig. 17.—The ratio r.m.s. rectifier current/average current per rectifier, plotted against $n\omega R_{LOAD}C$.
C in F, and R_{LOAD} in Ω.
n = 1 for half-wave
n = 2 for full-wave
n = 0·5 for voltage-doubler

(10) Check the peak recurrent current per rectifier from Fig. 18.

(11) Check the initial switch-on current I_{on} given by $E_{T(max)}/R_S$. If the value obtained exceeds that specified for the rectifier, then R_s must be increased and the design procedure repeated.

(12) Design the transformer and adjust the value of R_s accordingly, taking into account the transformer resistance and the forward resistance of the rectifier at the average current.

(13) Check the r.m.s. ripple current through the capacitor.

(14) Design the R-C damping circuit as recommended in the published data of the rectifier.

(15) Determine the size of heatsink to be used to allow operation at the desired ambient temperature (from published data).

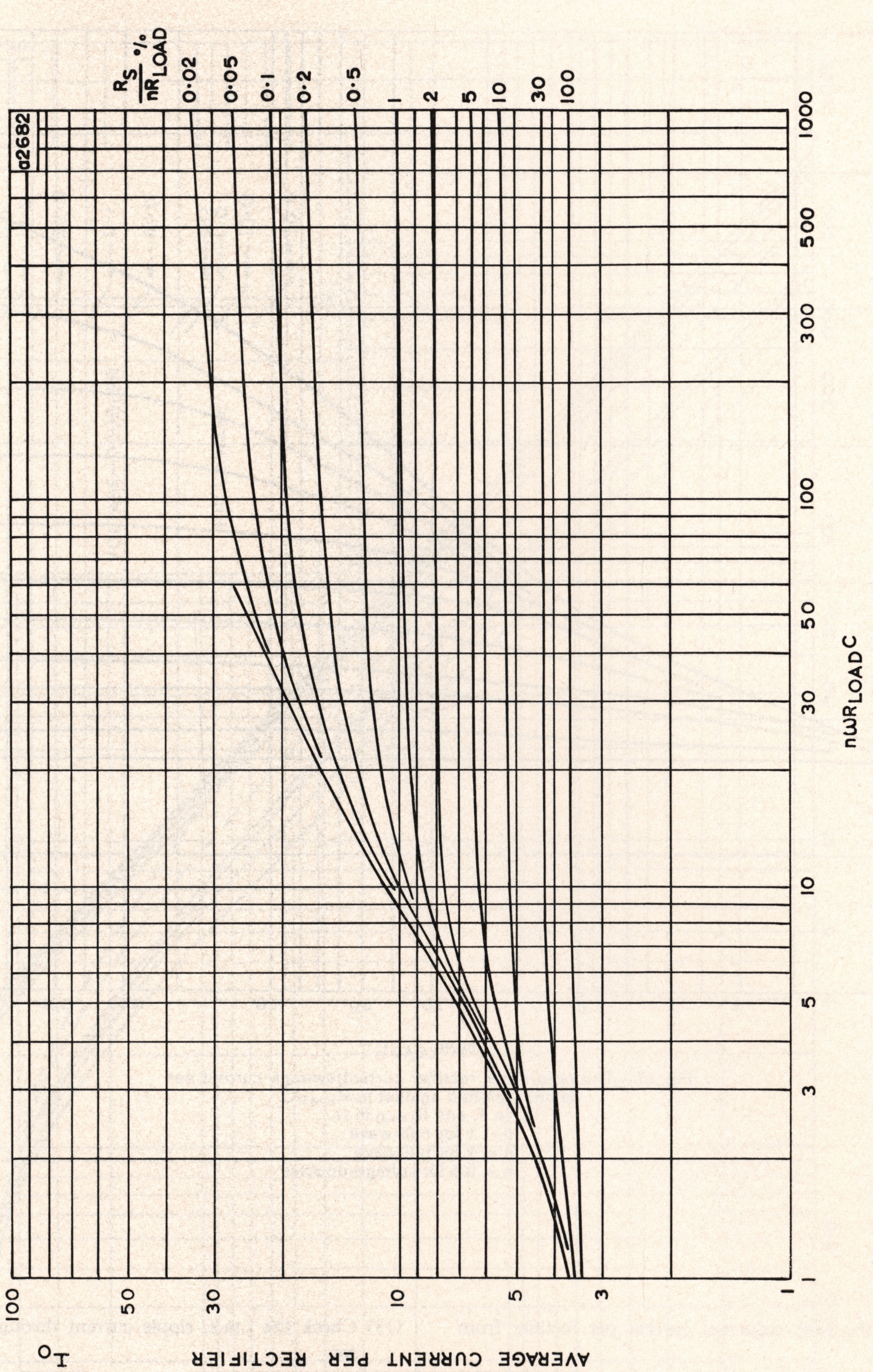


Fig. 18—The ratio recurrent peak current/average current per rectifier, plotted against $n\omega R_{LOAD}C$.
C in F, and R_{LOAD} in Ω. $\omega = 2\pi f$,
f = line frequency
n = 1 for half-wave
n = 2 for full-wave
n = 0·5 for voltage-doubler

Design Examples

The design of the four types of capacitor input filter rectifier circuit is illustrated in Table 4 on pages 244 and 245, the recommended procedure being used.

R-C Damping Circuit

The R-C damping circuit for the above four examples can be designed by adopting the following procedure. It may be connected to either the primary or the secondary of the transformer (see BYZ10 data).

Consider the full-wave bridge rectifier circuit. If the damping circuit is connected to the primary of the transformer, then

$$C_1 = 200\frac{I_{mag}}{V}\mu F \text{ and } R_1 = \frac{150}{C_1}\Omega$$

where

V = transformer primary r.m.s. voltage
I_{mag} = magnetising primary r.m.s. current (A)

From Table 4, the primary r.m.s. current = 3·68/0·892 = 4·13A.

If I_{mag} = 10% of primary r.m.s. current, then

$$C_1 = 200(0{\cdot}413/230) = 0{\cdot}36\mu F.$$

Let

$$C_1 = 0{\cdot}5\mu F$$

then

$$R_1 = 150/0{\cdot}5 = 300\Omega.$$

If the damping circuit is connected to the secondary,

$$C_2 = \frac{225(I_{mag}T^2)}{V}\mu F \text{ and } R_2 = \frac{200}{C_2}\Omega,$$

where

$$T = \frac{\text{transformer primary r.m.s. voltage}}{\text{transformer secondary r.m.s. voltage}}.$$

Therefore

$$C_2 = \frac{225 \times 0{\cdot}413}{230}\cdot\left(\frac{230}{258}\right)^2 = 0{\cdot}31\mu F.$$

Let

$$C_2 = 0{\cdot}5\mu F, \text{ then } R_2 = 400\Omega.$$

Heatsink Design

The heatsinks can be designed for the above four circuits from the information provided in the rectifier data. The heatsink size for the bridge rectifier circuit is determined below to illustrate the procedure.

From the BYZ10 data, for 1A average current, a rectifier mounted on a heatsink with a thermal resistance of 14·4°C/W ($\theta_i+\theta_h$ = 15°C/W) can be operated up to an ambient temperature of 63°C. If these fins are stacked to produce the bridge rectifier assembly, then the area of the heatsink should be approximately 30cm² (one face). A heatsink measuring 6cm × 6cm should therefore be satisfactory.

Performance

The voltage regulation curves for the four examples are shown in Figs. 19 to 22. From these curves it can be seen that the output voltage at the required current is within 2% of the stated value.

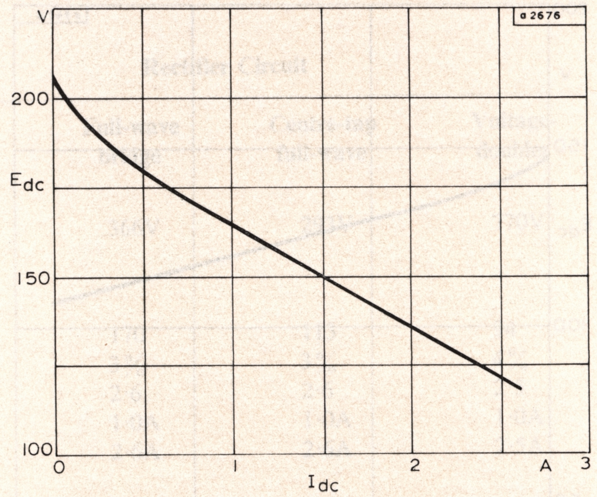


Fig. 19—Voltage regulation for single-phase half-wave circuit with capacitor input filter

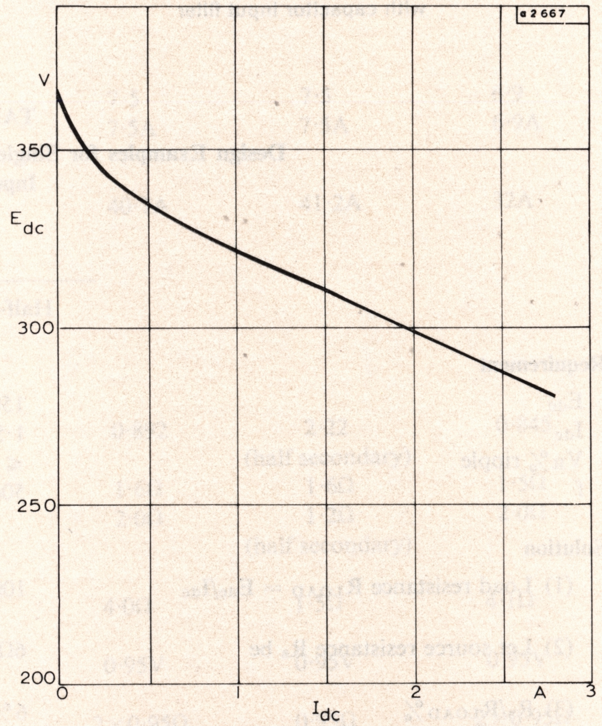


Fig. 20—Voltage regulation for single-phase full-wave bridge circuit with capacitor input filter

DESIGN PROCEDURE FOR RECTIFIER CIRCUITS WITH CHOKE INPUT FILTER

The analysis of the capacitor input filter rectifier circuits has shown that for any high-current conversion, the circuit requires a large value of smoothing capacitor, which has to carry a large ripple current; and large initial and recurrent peak currents flow through the rectifiers. These limitations may be overcome by the use of choke input filters.

The single-phase half-wave circuit (Fig. 4) cannot be used with a choke input filter, as it would require an infinite value of inductance to cause current to flow throughout the cycle.

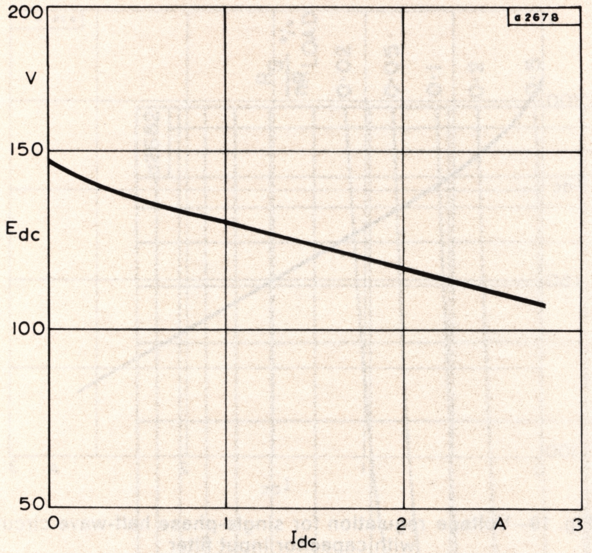


Fig. 21—Voltage regulation for single-phase centre-tap circuit with capacitor input filter

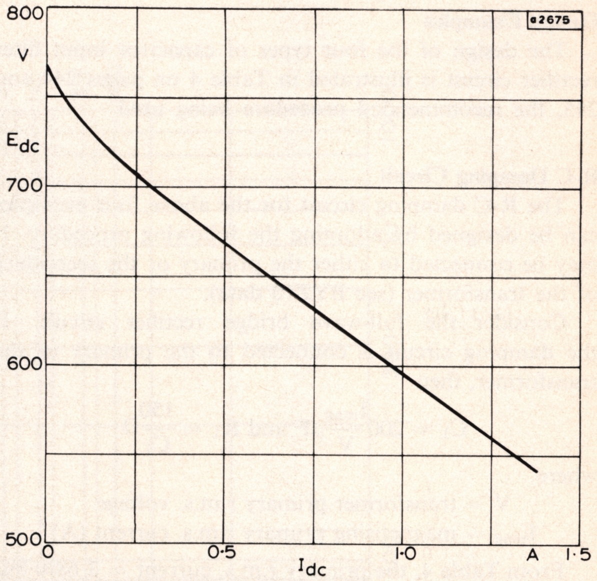


Fig. 22—Voltage regulation for single-phase voltage-doubler circuit with capacitor input filter

TABLE 4

Design Examples for Single-phase Circuits with Capacitor Input Filter

	Rectifier Circuit			
	Half-wave	Full-wave bridge	Centre-tap full-wave	Voltage-doubler
Requirement				
E_{dc}	150V	300V	120V	600V
I_{dc}	1·5A	2·0A	2·0A	1·0A
V_R% ripple	≤1%	≤1%	≤1·3%	≤1%
f	50c/s	50c/s	50c/s	50c/s
Solution				
(1) Load resistance $R_{LOAD} = E_{dc}/I_{dc}$	100Ω	150Ω	60Ω	600Ω
(2) Let source resistance R_s be	6Ω	9Ω	3·6Ω	12Ω
(3) R_s/R_{LOAD}%	6%	6%	6%	2%
(4) Value of $\omega R_{LOAD}C$ from Fig. 16	150	66	50	≃150
$C = \frac{(\omega R_{LOAD}C)}{2\pi 50 R_{LOAD}}$	4780μF	1400μF	2660μF	795μF
Practical value of C	5000μF	1800μF	3000μF	1000μF
New value of $\omega R_{LOAD}C$	157	85	56·5	188
(5) Conversion ratio $E_{dc}/E_{T(max)}$ using %R_s/R_{LOAD} from (3) and the new value of $\omega R_{LOAD}C$ from (4)	From Fig. 13 0·73	From Fig. 14 0·82	From Fig. 14 0·82	From Fig. 15 1·56
(6) $E_{T(max)} = \frac{E_{dc}}{\text{Conversion ratio}}$	205V	366V	146·5V	385V
$E_{rms} = \frac{E_{T(max)}}{\sqrt{2}}$	145V	258V	103·5V	272V

TABLE 4

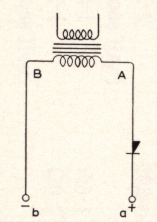
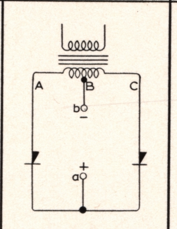
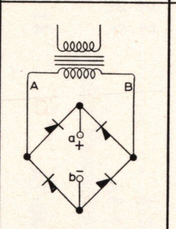
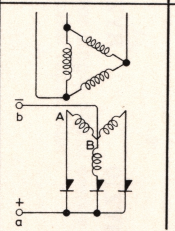
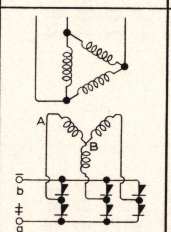
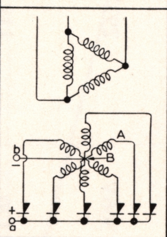
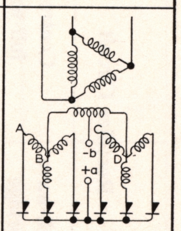
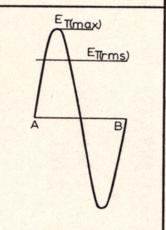
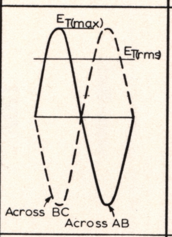
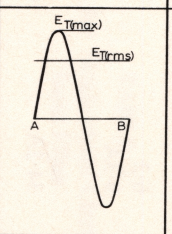
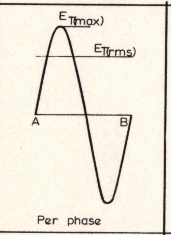
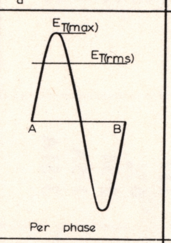
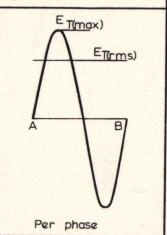
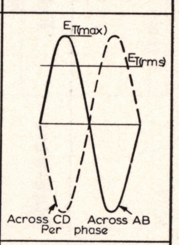
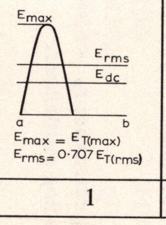
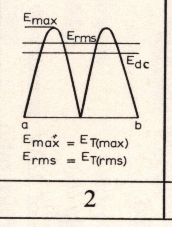
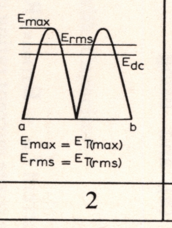
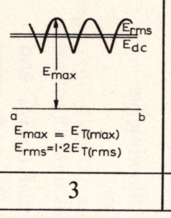
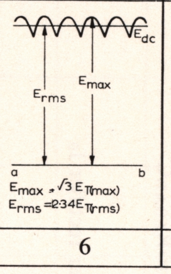
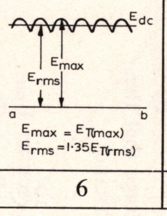
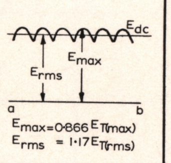
Continued

	Rectifier Circuit			
	Half-wave	Full-wave bridge	Centre-tap full-wave	Voltage-doubler
(7) Crest working voltage that the rectifiers must withstand	205V	366V	293V	770V
(8) r.m.s. current per rectifier I_{rms}, from Fig. 17, for $n\omega R_{LOAD}C$	157	170	113	94
R_s/nR_{LOAD}%	6%	3%	3%	4%
I_{rms}/I_0	2·34	2·6	2·6	2·5
Average current per rectifier I_0	1·5A	1·0A	1·0A	1·0A
therefore I_{rms}	3·51A	2·6A	2·6A	2·5A
(9) Suitable rectifiers, allowing for transient derating	BYZ12	BYZ10	BYZ11	2×BYZ10 in series
(10) Taking values of $n\omega R_{LOAD}C$ and R_s/nR_{LOAD}% in (8) and using Fig. 18, I_{pk}/I_0	6·2	7·5	7·5	6·9
therefore I_{pk}	9·3A	7·5A	7·5A	6·9A
(11) Initial switch-on current $I_{on} = E_{T(max)}/R_s$ Check this with the surge current rating of the rectifiers selected.	34·2A	40·7A	41·2A	32A
(12) Transformer design For mains voltage of 230V, primary to secondary transformer ratio $N = 230/E_{T(rms)}$	1·585	0·892	2·22 (half secondary)	0·845
If primary winding resistance r_p =	1·5Ω	1·6Ω	1·6Ω	1·5Ω
and secondary winding resistance r_s =	2·0Ω	2·0Ω	1·2Ω (half secondary)	2·0Ω
then transformer resistance referred to secondary is $r_s + (r_p/N^2)$ =	2·6Ω	4·0Ω	1·5Ω	4·1Ω
Voltage drop V_D across rectifier at average current I_0 =	1·0V	0·95V	0·95V	0·95V
therefore rectifier resistance in circuit at average current is $r_r = V_D/I_0$ =	0·67Ω	2×0·95Ω	0·95Ω	2×0·95Ω
Total resistance in secondary circuit $= r_s + (r_p/N^2) + r_r = r_{tot}$	3·27Ω	5·9Ω	2·45Ω	6·0Ω
External series resistance must be $R_s - r_{tot}$ =	2·73Ω	3·1Ω	1·15Ω	6·0Ω
Let $R_s - r_{tot}$ be	3·0Ω	3·0Ω	1·0Ω	6·0Ω
Secondary r.m.s. current $I_{T(rms)}$	3·51A	√2×2·6 = 3·68A	2·6A (half secondary)	√2×2·5 = 3·55A
Secondary r.m.s. voltage $E_{T(rms)}$	145V	258V	103·5V	272V
Secondary Volt-Amp rating $VA_s = E_{T(rms)} . I_{T(rms)}$	508VA	950VA	269+269VA	963VA
Power rating of series resistor	37W	40·6W	13·6W	75W
(13) r.m.s. ripple current $I_{c(rms)}$	3·18A (Eq 2)	3·54A (Eq 3)	3·09A (Eq 3)	2·29A (Eq 2)

See REVISED VERSION - July 1966, pp. 46-47.

TABLE 1

IDEALISED RECTIFIER CIRCUIT PERFORMANCES

	Single-Phase			Three-Phase			
	Half-Wave	**Centre-Tap Full-Wave**	**Full-Wave Bridge**	**Half-Wave**	**Full-Wave Bridge**	**Centre-Tap**	**Double-Star**
Type of rectifier circuit							
Secondary input voltage per phase							
Output voltage across a-b	 $E_{max} = E_{T(max)}$ $E_{rms} = 0{\cdot}707E_{T(rms)}$	 $E_{max} = E_{T(max)}$ $E_{rms} = E_{T(rms)}$	 $E_{max} = E_{T(max)}$ $E_{rms} = E_{T(rms)}$	 $E_{max} = E_{T(max)}$ $E_{rms} = 1{\cdot}2E_{T(rms)}$	 $E_{max} = \sqrt{3}E_{T(max)}$ $E_{rms} = 2{\cdot}34E_{T(rms)}$	 $E_{max} = E_{T(max)}$ $E_{rms} = 1{\cdot}35E_{T(rms)}$	 $E_{max} = 0{\cdot}866E_{T(max)}$ $E_{rms} = 1{\cdot}17E_{T(rms)}$
Number of Output Voltage Pulses per Cycle (N)	1	2	2	3	6	6	6
OUTPUT VOLTAGE							
E_{dc} in terms of r.m.s. Input Voltage per Phase $E_{T(rms)}$	$0{\cdot}45E_{T(rms)}$	$0{\cdot}90E_{T(rms)}$	$0{\cdot}90E_{T(rms)}$	$1{\cdot}17E_{T(rms)}$	$2{\cdot}34E_{T(rms)}$	$1{\cdot}35E_{T(rms)}$	$1{\cdot}17E_{T(rms)}$
E_{dc} in terms of r.m.s. Output Voltage E_{rms}	$0{\cdot}636E_{rms}$	$0{\cdot}90E_{rms}$	$0{\cdot}90E_{rms}$	$0{\cdot}98E_{rms}$	E_{rms}	E_{rms}	E_{rms}
E_{dc} in terms of Peak Output Voltage E_{max}	$0{\cdot}318E_{max}$	$0{\cdot}636E_{max}$	$0{\cdot}636E_{max}$	$0{\cdot}826E_{max}$	$0{\cdot}955E_{max}$	$0{\cdot}955E_{max}$	$0{\cdot}955E_{max}$
R.M.S. Output Voltage E_{rms} in terms of E_{dc}	$1{\cdot}57E_{dc}$	$1{\cdot}11E_{dc}$	$1{\cdot}11E_{dc}$	$1{\cdot}02E_{dc}$	$1{\cdot}00E_{dc}$	$1{\cdot}00E_{dc}$	$1{\cdot}00E_{dc}$
Peak Output Voltage E_{max} in terms of E_{dc}	$3{\cdot}14E_{dc}$	$1{\cdot}57E_{dc}$	$1{\cdot}57E_{dc}$	$1{\cdot}21\ E_{dc}$	$1{\cdot}05E_{dc}$	$1{\cdot}05E_{dc}$	$1{\cdot}05E_{dc}$

OUTPUT CURRENT								
Average Current per Rectifier Leg I_o		I_{dc}	$0{\cdot}5I_{dc}$	$0{\cdot}5I_{dc}$	$0{\cdot}33I_{dc}$	$0{\cdot}33I_{dc}$	$0{\cdot}167I_{dc}$	$0{\cdot}167I_{dc}$
I_{rms} per Rectifier Leg	R	$1{\cdot}57I_{dc}$	$0{\cdot}785I_{dc}$	$0{\cdot}785I_{dc}$	$0{\cdot}588I_{dc}$	$0{\cdot}577I_{dc}$	$0{\cdot}408I_{dc}$	$0{\cdot}293I_{dc}$
	L		$0{\cdot}707I_{dc}$	$0{\cdot}707I_{dc}$	$0{\cdot}577I_{dc}$	$0{\cdot}577I_{dc}$	$0{\cdot}408I_{dc}$	$0{\cdot}289I_{dc}$
I_{pk} per Rectifier Leg	R	$3{\cdot}14I_{dc}$	$1{\cdot}57I_{dc}$	$1{\cdot}57I_{dc}$	$1{\cdot}21I_{dc}$	$1{\cdot}05I_{dc}$	$1{\cdot}05I_{dc}$	$0{\cdot}525I_{dc}$
	L		I_{dc}	I_{dc}	I_{dc}	I_{dc}	I_{dc}	$0{\cdot}5I_{dc}$
TRANSFORMER RATING								
Secondary r.m.s. Voltage per Transformer Leg $E_{T(rms)}$		$2{\cdot}22E_{dc}$	$1{\cdot}11E_{dc}$ (to centre-tap)	$1{\cdot}11E_{dc}$ (total)	$0{\cdot}855E_{dc}$ (to neutral)	$0{\cdot}428E_{dc}$ (to neutral)	$0{\cdot}74E_{dc}$ (to neutral)	$0{\cdot}855E_{dc}$ (to neutral)
Secondary r.m.s. Current per Transformer Leg $I_{T(rms)}$	R	$1{\cdot}57I_{dc}$	$0{\cdot}785I_{dc}$	$1{\cdot}11I_{dc}$	$0{\cdot}588I_{dc}$	$0{\cdot}816I_{dc}$	$0{\cdot}408I_{dc}$	$0{\cdot}293I_{dc}$
	L		$0{\cdot}707I_{dc}$	I_{dc}	$0{\cdot}577I_{dc}$	$0{\cdot}816I_{dc}$	$0{\cdot}408I_{dc}$	$0{\cdot}289I_{dc}$
Secondary Volt-Amp VA_s	R	$3{\cdot}48E_{dc}.I_{dc}$	$1{\cdot}74E_{dc}.I_{dc}$	$1{\cdot}23E_{dc}.I_{dc}$	$1{\cdot}50E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$	$1{\cdot}81E_{dc}.I_{dc}$	$1{\cdot}50E_{dc}.I_{dc}$
	L		$1{\cdot}57E_{dc}.I_{dc}$	$1{\cdot}11E_{dc}.I_{dc}$	$1{\cdot}48E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$	$1{\cdot}81E_{dc}.I_{dc}$	$1{\cdot}48E_{dc}.I_{dc}$
Secondary Utility Factor U_s	R	0·287	0·574	0·813	0·666	0·95	0·552	0·666
	L		0·636	0·90	0·675	0·95	0·552	0·675
Primary Voltage per Transformer Leg (Transformer Ratio 1:1)		$2{\cdot}22E_{dc}$	$1{\cdot}11E_{dc}$	$1{\cdot}11E_{dc}$	$0{\cdot}855E_{dc}$	$0{\cdot}428E_{dc}$	$0{\cdot}74E_{dc}$	$0{\cdot}855E_{dc}$
Primary Current per Transformer Leg (Transformer Ratio 1:1)	R	$1{\cdot}57I_{dc}$	$1{\cdot}11I_{dc}$	$1{\cdot}11I_{dc}$	$0{\cdot}588I_{dc}$	$0{\cdot}816I_{dc}$	$0{\cdot}577I_{dc}$	$0{\cdot}408I_{dc}$
	L		I_{dc}	I_{dc}	$0{\cdot}471I_{dc}$	$0{\cdot}816I_{dc}$	$0{\cdot}577I_{dc}$	$0{\cdot}408I_{dc}$
Primary Volt-Amp VA_p	R	$3{\cdot}48E_{dc}.I_{dc}$	$1{\cdot}23E_{dc}.I_{dc}$	$1{\cdot}23E_{dc}.I_{dc}$	$1{\cdot}50E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$	$1{\cdot}28E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$
	L		$1{\cdot}11E_{dc}.I_{dc}$	$1{\cdot}11E_{dc}.I_{dc}$	$1{\cdot}21E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$	$1{\cdot}28E_{dc}.I_{dc}$	$1{\cdot}05E_{dc}.I_{dc}$
Primary Utility Factor U_p	R	0·287	0·813	0·813	0·666	0·95	0·78	0·95
	L		0·90	0·90	0·827	0·95	0·78	0·95
Fundamental Ripple Frequency f_r		f	2f	2f	3f	6f	6f	6f
$\% \text{ Ripple} = \dfrac{\text{r.m.s. Fundamental Ripple Voltage}}{E_{dc}} \times 100$		111	47·2	47·2	17·7	4·0	4·0	4·0
Crest Working Voltage In terms of E_{dc}		$3{\cdot}14E_{dc}$	$3{\cdot}14E_{dc}$	$1{\cdot}57E_{dc}$	$2{\cdot}09E_{dc}$	$1{\cdot}05E_{dc}$	$2{\cdot}09E_{dc}$	$2{\cdot}42E_{dc}$
In terms of $E_{T(rms)}$		$1{\cdot}41E_{T(rms)}$	$2{\cdot}82E_{T(rms)}$	$1{\cdot}41E_{T(rms)}$	$2{\cdot}45E_{T(rms)}$	$2{\cdot}45E_{T(rms)}$	$2{\cdot}83E_{T(rms)}$	$2{\cdot}83E_{T(rms)}$

R = Resistive Load L = Inductive Load f = Supply Frequency c/s.

In the calculation of the above circuit performances, the rectifier forward voltage drop and the transformer impedance have been ignored. The primary volt-amp rating of the transformer does not take primary magnetising current into account.

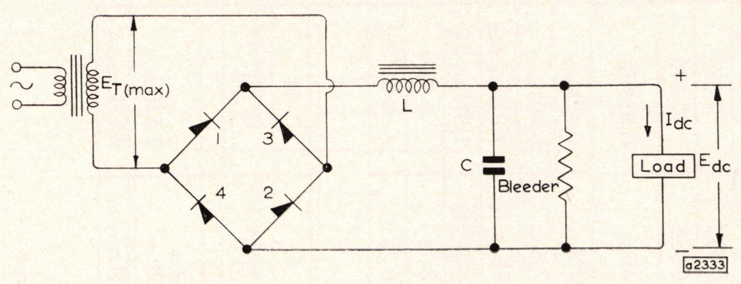


Fig. 23—Single-phase full-wave bridge circuit with choke input filter

For the full-wave bridge circuit of Fig. 6 and the full-wave centre-tap circuit of Fig. 7, R_s is replaced by a series choke L.

The full-wave bridge circuit with choke input filter is shown in Fig. 23, and the voltage and current waveforms in Fig. 24. The action of the choke is to reduce both the peak and the r.m.s. value of current and to reduce the ripple voltage. The choke input filter circuit, however, requires a higher applied voltage than the capacitor input filter circuit, to produce the same output voltage.

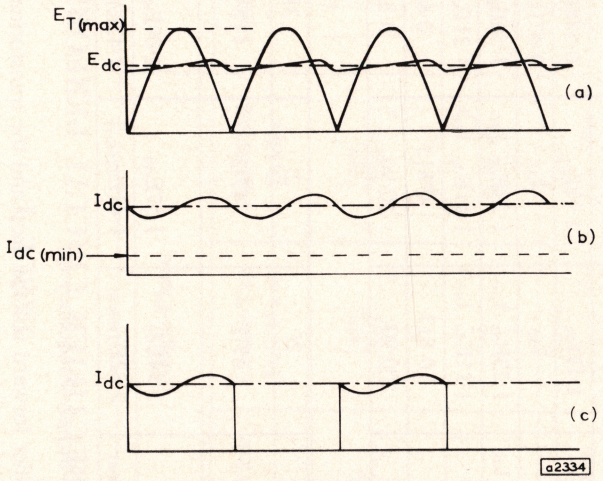


Fig. 24—Waveforms for full-wave bridge circuit with choke input filter
(a) output voltage
(b) current through choke
(c) current through rectifiers (1 and 2) or (3 and 4)

Smoothing Circuit

The choke input filter must, ideally, pass only one frequency, which is zero, and attenuate all others. The filter must allow direct current to flow to the load without much power loss, and at the same time present a high impedance to the fundamental and other ripple frequencies. The capacitor shunts the load so as to by-pass the harmonic currents.

The attenuation factor K of the filter with series choke L and shunt capacitor C, is defined as the ratio of the total input impedance of the filter to the impedance of the parallel combination of the shunt capacitor C and load R_{LOAD}. For the choke input filter to function efficiently, the choke reactance at fundamental ripple frequency f_r should be much greater than its d.c. resistance, and the capacitor reactance much lower than the minimum load resistance.

If it is assumed that, the inductance of the choke being L,

$$2\pi f_r L \gg \text{choke resistance } R_L\,,$$

and

$$\frac{1}{2\pi f_r C} \ll R_{LOAD(min)}$$

then

$$K = \frac{2\pi f_r L - \dfrac{1}{2\pi f_r C}}{\dfrac{1}{2\pi f_r C}}$$

therefore

$$K = 4\pi^2 f_r^2 LC - 1. \qquad \ldots(4)$$

The value of the inductance L used in the circuit must be such as to allow the rectifiers to conduct over one cycle of the fundamental ripple frequency. If the rectifier conducts for a period less than this, then the choke input filter will behave more and more like a capacitor input filter. This will give rise to a higher recurrent peak current through the rectifiers, and will also result in poor regulation.

The use of sufficient inductance allows the rectifier to conduct over the complete cycle; whereas the capacitor input filter allows the rectifier to conduct over only a fraction of a cycle. It follows that, for a given current, the rectifier will switch off before the cycle is completed, for a certain value of inductance. This value is termed the critical inductance L_{crit}.

Output Voltage

Consider the single-phase full-wave bridge circuit shown in Fig. 23 and the waveforms shown in Fig. 24. The rectified voltage is applied to the choke input filter. This voltage may be expressed as a series containing a d.c. component and harmonic components, as shown on page 232. The crest value of the output voltage is E_{max}, and it is equal to $E_{T(max)}$ in this circuit.

The rectified voltage can be approximated to a d.c. term plus a harmonic at the fundamental ripple frequency, assuming that the amplitudes of the higher harmonics are negligible. Eq (1) is derived on this assumption. Therefore, as derived on page 232,

$$e \simeq \frac{2}{\pi}E_{max} - \frac{4}{3\pi}E_{max}\cos 2\omega t.$$

Critical Inductance

From Fig. 24 it can be seen that for the rectifier to conduct throughout the fundamental ripple cycle, the negative-going peak ripple current delivered by the rectifier must not exceed the minimum d.c. current, which occurs with a load of $R_{LOAD(max)}$. Thus

$$I_{dc(min)} = \frac{E_{dc}}{R_{LOAD(max)}} = \frac{2E_{max}}{\pi}\frac{1}{R_{LOAD(max)}}. \qquad \ldots(5)$$

If $2\pi f_r L \gg R_L$, and

$$\frac{1}{2\pi f_r C} \ll R_{LOAD(min)}$$

$$\text{peak a.c. current} = \frac{4}{3\pi}E_{max}\frac{1}{2\pi f_r L}. \qquad \ldots(6)$$

The critical inductance is reached when the peak a.c. current equals the direct current. That is,

$$\frac{4}{3\pi}E_{max}\frac{1}{2\pi f_r L_{crit}} = \frac{2E_{max}}{\pi}\cdot\frac{1}{R_{LOAD(max)}}$$

therefore

$$L_{crit} = \frac{R_{LOAD(max)}}{3\pi f_r}. \qquad \ldots(7)$$

For 50c/s supply frequency and full-wave rectification, f_r = 100c/s, so that

$$L_{crit} = \frac{R_{LOAD(max)}}{943}. \qquad \ldots(8)$$

Because of the approximations made, it is necessary to use a somewhat higher value of inductance than L_{crit}. In practice, it is found that for reliable and satisfactory operation the optimum value of inductance that should be used is twice the value of L_{crit}.

It is obvious from the nature of the circuit that it is not possible to maintain the critical value of the inductance over all values of load current. This would require an infinite inductance at zero load current. Two methods are available to ensure that current flows throughout the cycle, and that good regulation is maintained over a wide range of load currents. These are the use of a bleeder resistance or a swinging choke.

Bleeder Resistance

A bleeder resistance of a suitable value is connected across the shunt capacitor to maintain the minimum current that will satisfy the critical inductance condition, even when no load is connected. The use of a bleeder will prevent the output voltage from rising to the peak applied voltage in the absence of the load.

Swinging Choke

The swinging choke method is based on the fact that the inductance of an iron-cored inductor partly depends on the amount of direct current flowing through it. The swinging choke is designed so that it has a high inductance value at low currents, and this decreases as the d.c. current is increased. The use of such a choke is therefore very satisfactory for maintaining good regulation over a range of load current; and it is also more efficient than the bleeder resistance method.

Since the inductance is continually varying with the load current, the ripple voltage is no longer independent of the load current. When using a swinging choke, it is necessary to ensure that the inductance does not fall to a very low value at the maximum load current, as this will lead to high recurrent peak currents. In practice, the inductance at full load L_F should be such that

$$L_F = 2R_{LOAD(min)}/943.$$

Ripple Current and Voltage

If $2\pi f_r L \gg R_L$,

$$\frac{1}{2\pi f_r C} \ll R_{LOAD(min)}$$

and

$$2\pi f_r L \gg \frac{1}{2\pi f_r C}.$$

then

$$\text{r.m.s. ripple current } I_{c(rms)} = \left(\frac{4}{3}\cdot\frac{E_{max}}{\pi}\cdot\frac{1}{\sqrt{2}}\right)\frac{1}{2\pi f_r L}.$$

Since

$$E_{dc} = \frac{2}{\pi}E_{max}\,,$$

$$I_{c(rms)} = \frac{\sqrt{2}}{3}E_{dc}\frac{1}{2\pi f_r L} \qquad \ldots(9)$$

$$\%\ \text{ripple} = \%\ \text{ripple before filtering} \times 1/K.$$

From Table 1, % ripple before filtering = 47·2%.

From Eq(4), if $4\pi^2 f_r^2 LC \gg 1$

then

$$K \simeq 4\pi^2 f_r^2 LC$$

$$\%\ \text{ripple} = \frac{47{\cdot}2}{4\pi^2 f_r^2 LC} = \frac{1{\cdot}193}{f_r^2 LC}. \qquad \ldots(10)$$

For 50c/s supply frequency and full-wave rectification, f_r = 100c/s, therefore

$$\%\ \text{ripple} = 119{\cdot}3/LC, \qquad \ldots(11)$$

where L is in henries and C is in μF.

Minimum Value of Shunt Capacitance

In evaluating the percentage ripple and the attenuation factor of the filter, it has been assumed that the reactance of the capacitor at the fundamental ripple frequency is very much lower than the minimum load resistance. In practice, it is found that satisfactory performance is obtained when the reactance of the capacitor is made less than one-fifth the minimum load resistance. That is,

$$\frac{1}{2\pi f_r C} \leqslant R_{LOAD(min)}/5.$$

Therefore

$$C \geqslant \frac{5\times 10^6}{2\pi}\frac{1}{f_r R_{LOAD(min)}}\mu F$$

$$\geqslant \frac{796\,000}{f_r R_{LOAD(min)}}\mu F. \qquad \ldots(12)$$

Because of the nature of the circuit, the capacitor will resonate with the inductor at a certain frequency. At this frequency the output impedance will be greater than the capacitor reactance. Therefore, when a non-linear loading is applied, precautions must be taken to ensure that the output impedance of the filter is small at the load current frequency.

Additional Filter Sections

When it is required to reduce the ripple voltage across the load to a very low value, a single-stage choke input

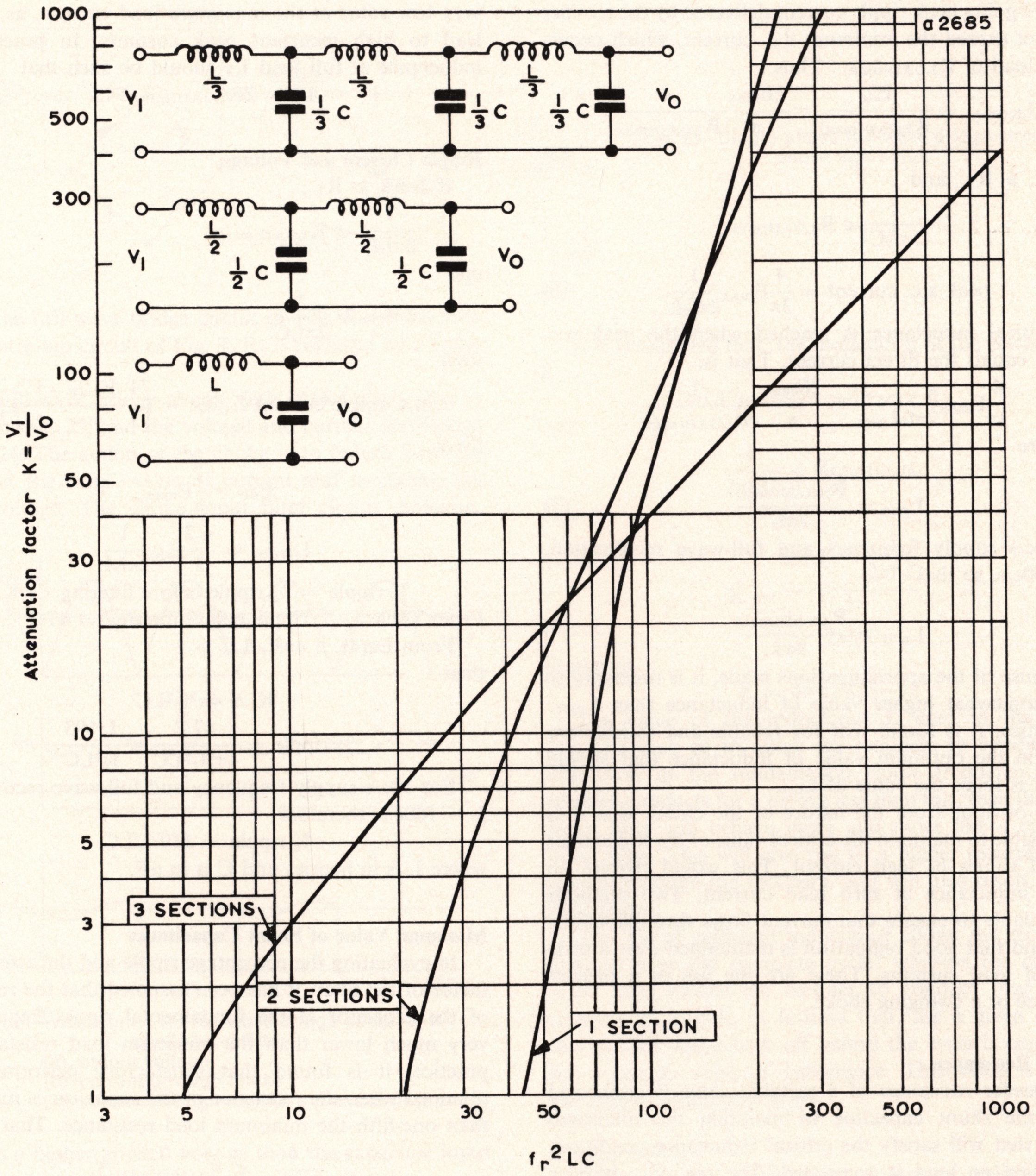


Fig. 25—Characteristics of choke input filters
L in H, and C in μF

filter may require large values of inductance and capacitance, which may lead to an uneconomic filter design. In this case, the same results may be achieved by using a multi-stage filter with small value inductors and capacitors. It can be shown that the optimum smoothing is achieved when all stages are identical.

Fig. 25 shows the attenuation factor K plotted against f_r^2LC for 1-, 2-, and 3-stage filters. A suitable arrangement can therefore be selected by studying the filter characteristics. For a K factor between 23 and 160, the two-stage filter is the most economic. For K above 160, a three-stage filter is more suitable.

Example on Full-wave Bridge Rectifier Circuit with Choke Input Filter

A full-wave bridge circuit to supply 0 to 4A at 200V is required. The ripple is to be less than 0·5%, f = 50c/s, choke resistance 7·5Ω, and rectifier voltage drop approximately 1V.

Let the bleeder current = 0·5A, therefore the bleeder resistance $R_b = 200/0{\cdot}5 = 400\Omega$.

The external load at maximum current $= R_{LOAD(min)} = 200/4 = 50\Omega$.

At zero load current, the total circuit resistance $\simeq 400+7{\cdot}5 = 407{\cdot}5\Omega$.

From Eq (8)

$$L_{crit} = R_{LOAD(max)}/943 = 407{\cdot}5/943 = 0{\cdot}432\text{H},$$

therefore the optimum value is 0·864H.

The relationships between a.c. and d.c. voltages and currents are given in Table 1 for circuits without filters. The values given for inductive load circuits can also be used for choke input filter circuits.

In order to use the relationship of the idealised circuits (Table 1), E_{dc} must be increased above the output direct voltage required to allow for voltage drop across the choke and rectifiers. That is,

$$E_{dc} = \text{direct output voltage required} + \text{volt drop across choke} + \text{volt drop across rectifiers.}$$

Therefore

$$E_{dc} = 200+7{\cdot}5(4+0{\cdot}5)+(2\times1) \simeq 236\text{V}.$$

From Table 1,

$$E_{rms} = 1{\cdot}11E_{dc} = 262\text{V}.$$

From Eq (12)

$$C \geqslant \frac{796\,000}{f_rR_{LOAD(min)}}.$$

For a bridge rectifier circuit, $f_r = 100$c/s, therefore

$$C \geqslant \frac{796\,000}{100\times50} \geqslant 160\mu\text{F}.$$

From Eq (11), for ripple to be less than 0·5%

$$LC \geqslant 119{\cdot}3/0{\cdot}5$$

therefore

$$LC \geqslant 238{\cdot}6.$$

If L = 1H, then $C \geqslant 238{\cdot}6\mu$F, so that a practical value of $C = 250\mu$F is suitable.

From Table 1, the crest working voltage that the rectifiers must withstand $= 1{\cdot}57E_{dc} = 1{\cdot}57\times236 = 370$V.

With allowance made for transients, BYZ10 rectifiers should operate satisfactorily in this circuit.

From Table 1, $I_{pk} = 1{\cdot}0I_{dc}$ for a pure inductive load. From Fig. 24 it can be seen that I_{pk} is greater than I_{dc}, but not as large as the peak current with a resistive or capacitive load.

The maximum recurrent peak current per rectifier $I_{pk} < 1{\cdot}57I_{dc}$, therefore

$$I_{pk} < 1{\cdot}57\times4{\cdot}5 < 7{\cdot}6\text{A}.$$

The transformer rating can be determined by a similar procedure to that shown for the capacitor input filter rectifier circuits (page 245). The transformer resistance must be taken into account, and the transformer ratio determined accordingly to give 262V r.m.s. on the secondary.

For a mains voltage of 230V, primary winding resistance $r_p = 1\Omega$, and secondary winding resistance $r_s = 1\Omega$, the transformer ratio is

$$N = \frac{V_p}{V_s} = \frac{230}{262+\left(r_s+\dfrac{r_p}{N^2}\right)I_{dc}} = 0{\cdot}843.$$

$$\text{Secondary volt-amp rating} = \frac{230}{0{\cdot}843}I_{dc} = 1230\text{VA}.$$

The R-C damping circuit and heatsink must be designed according to the procedure given for the capacitor input filter rectifier circuits (page 243).

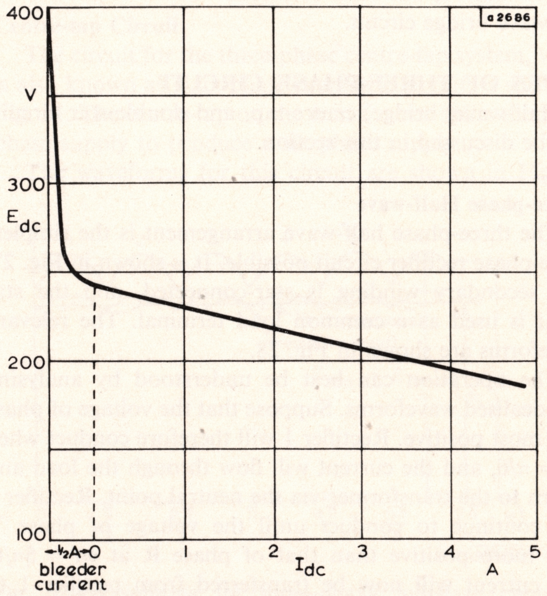


Fig. 26—Voltage regulation for single-phase, full-wave bridge circuit with choke input filter

The voltage regulation curve for a circuit built with these components is shown in Fig. 26, from which it can be seen that the output voltage at full load current is within 3% of the specified value. It can also be seen that the bleeder resistance is functioning correctly, since the current at which the voltage starts to rise rapidly is about one-half of the bleeder current.

Three-Phase Rectifier Circuits

GENERAL CONSIDERATIONS

There are many advantages in using a polyphase rectifier system when high-power conversion is required. The object is to superimpose more voltages of the same peak value but in different time relation to each other. An increase in the number of phases leads to the following improvements.

(i) Higher output voltage E_{dc} for the same voltage input.

(ii) Higher fundamental ripple frequency and lower amplitude ripple voltage.

(iii) Higher overall efficiency.

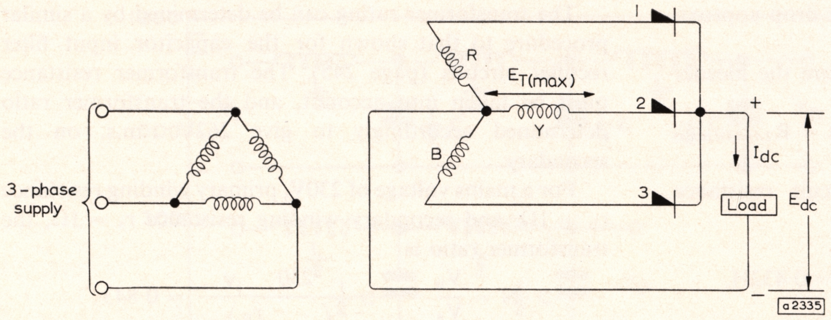


Fig. 27—Three-phase half-wave circuit

For three-phase circuits, one winding of the transformer is generally connected in delta to suppress harmonics (with the special exception of the second double-star circuit on page 254). In the explanation of the circuits in the next section, the secondary winding is always star-connected, but delta connection could be used in the full-wave bridge circuit.

TYPES OF THREE-PHASE CIRCUIT

Half-wave, bridge, centre-tap, and double-star circuits will be discussed in this section.

Three-phase Half-wave

The three-phase half-wave arrangement is the simplest three-phase rectifier circuit possible. It is shown in Fig. 27. The secondary winding is star-connected, and the star point is used as a common load terminal. The relevant waveforms are shown in Fig. 28.

The operation can best be understood by analysing the idealised waveforms. Suppose that the voltage of phase R is most positive. Rectifier 1 will therefore conduct when $\omega t = \pi/6$, and the current will flow through the load and return to the transformer via the neutral point. Rectifier 1 will continue to conduct until the voltage of phase Y goes more positive than that of phase R at $\omega t = 5\pi/6$. The current will now be transferred from rectifier 1 to rectifier 2. Rectifier 2 will conduct for the next 120°, and then the current will be transferred to rectifier 3 for the next 120°. In this manner, each rectifier conducts in turn for 120°.

The ripple frequency is three times the supply frequency; and the crest working voltage that the rectifiers must withstand is

$$2E_{T(\max)}\cos\frac{\pi}{6} = \sqrt{3}E_{T(\max)}.$$

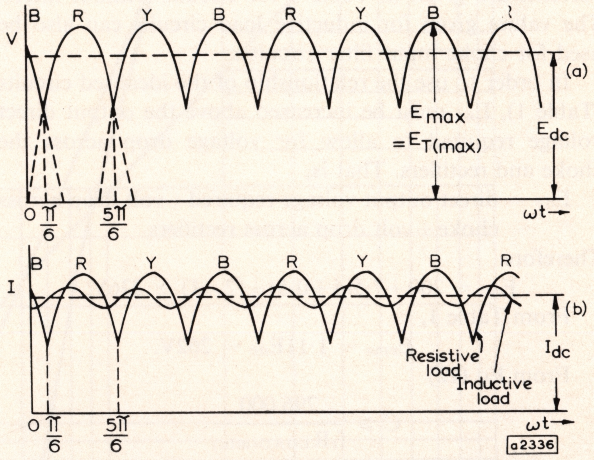


Fig. 28—Waveforms for three-phase half-wave circuit
(a) voltage (b) current

The conversion efficiency of this circuit is high in comparison with single-phase circuits, and the ripple voltage is reduced to little more than one-third of that obtained in the single-phase full-wave circuit. The transformer utility factor is, however, poor in comparison to the three-phase full-wave bridge rectifier, and the circuit is used only where low-voltage conversion is required.

Full-wave Bridge Circuit

The three-phase full-wave bridge circuit is shown in Fig. 29. It is one of the most widely-used circuits for high-power conversion with semiconductor rectifiers.

Consider the circuit in conjunction with the waveforms shown in Fig. 30. If phase R is most positive, rectifier 1 will start conducting when $\omega t = \pi/6$. The current flows through rectifier 1 to the load and returns to the transformer through rectifier 5 or 6, depending on which phase, Y or B, is the more negative. At $\omega t = \pi/6$, phase Y is the most negative and therefore current will flow through rectifier 5. At $\omega t = \pi/2$ phase B goes more negative, and therefore current will now flow through rectifier 6 instead of rectifier 5. At $\omega t = 5\pi/6$, phase Y goes more positive and the current is therefore transferred from rectifier 1 to rectifier 2. Each rectifier conducts for 120° per cycle, and the current is commutated every 60°.

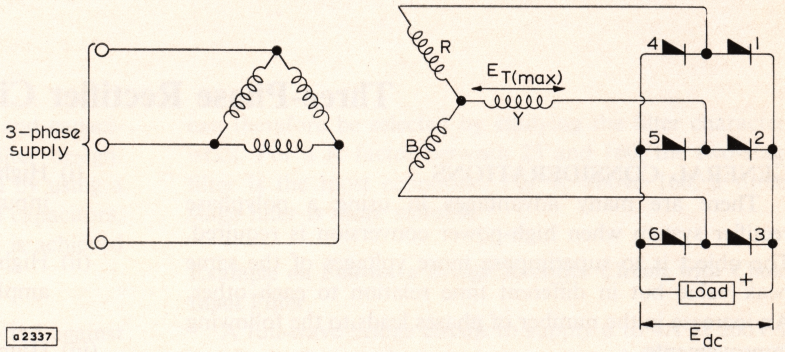


Fig. 29—Three-phase full-wave bridge circuit

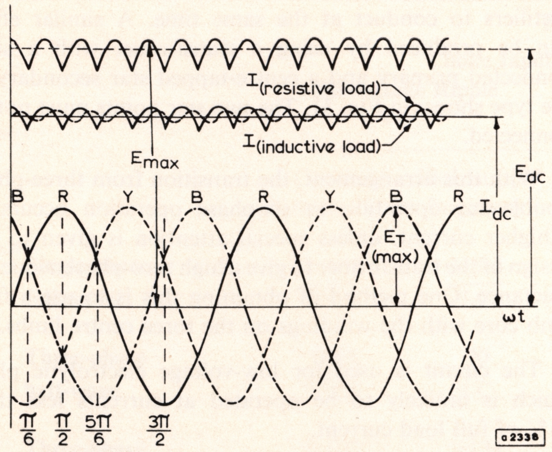


Fig. 30—Waveforms for three-phase full-wave bridge circuit

As in the single-phase bridge circuit (Fig. 6) the crest working voltage, as given in Table 1, appears across two rectifiers. The ripple voltage is small, and the ripple frequency is six times the supply frequency. This circuit has the highest transformer utility factor, and it therefore requires least a.c. power to obtain a specified direct voltage and current.

The circuit finds its applications in the charging of higher-voltage batteries, industrial power supplies, electrolytic plant operating at any voltage except very low voltages, and generally where high-power conversion is required most efficiently and economically.

Double Bridge Circuit

The double bridge circuit may be used where a very low ripple voltage is required. The primary winding is either delta- or star-connected. There are two sets of secondary windings. One set is star-connected and the other is delta-connected. Each set of windings is connected to a three-phase full-wave bridge rectifier assembly (the Fig. 29 circuit), and the output terminals of the two bridge circuits are connected in parallel. If a three-wire (centre earth) d.c. output is required, then the output terminals are connected in series.

The phase voltage of the secondary delta winding is $\sqrt{3}$ times the phase voltage of the secondary star winding; so that the amplitudes of the output voltages from both the bridge rectifier circuits are the same. However, the output voltage from the delta circuit is displaced in phase by $\pi/6$ relative to the output voltage from the star circuit. The ripple frequency is therefore twelve times the supply frequency. The percentage ripple is approximately 0·985%, and the output voltage $E_{dc} = 0{\cdot}99E_{max}$ or $1{\cdot}71E_{T(max)}$.

Centre-tap Circuit

The circuit for the three-phase centre-tap system, which is also known as a six-phase diametric circuit, is shown in Fig. 31. The centre-tap on the transformer splits the three-phase supply to produce a six-phase supply.

The waveforms for this circuit are shown in Fig. 32.

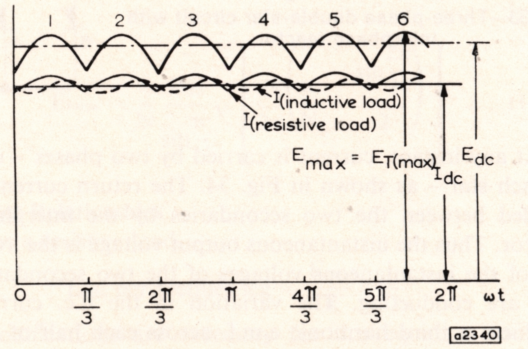


Fig. 32—Waveforms for three-phase full-wave centre-tap circuit

Each rectifier conducts for 60°, and the ripple frequency is six times the supply frequency. This system has higher conversion efficiency than the half-wave three-phase circuit. However, it has the lowest secondary utility factor of any

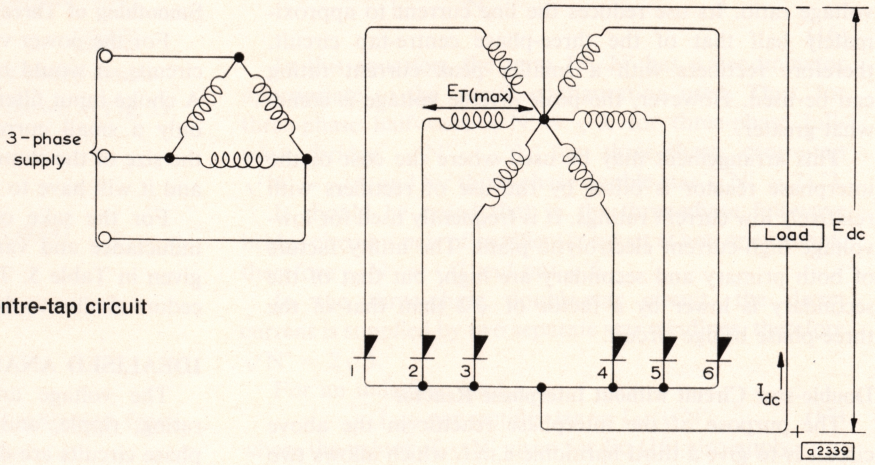


Fig. 31—Three-phase full-wave centre-tap circuit

three-phase circuit. The conversion efficiency is high, and equal to that of the three-phase bridge.

The chief attraction of the circuit is that all rectifiers are connected to a common terminal, and therefore can be simply mounted on one heatsink. It is generally used for low-power conversion only, because of the poor secondary utility factor.

Double-star Circuit with Interphase Reactor

The double-star circuit with interphase reactor is shown in Fig. 33. It has, in effect, two star-connected secondaries. The voltages of the two star connections are displaced by 180°. The neutral points of the two windings are connected together by a centre-tapped interphase reactor.

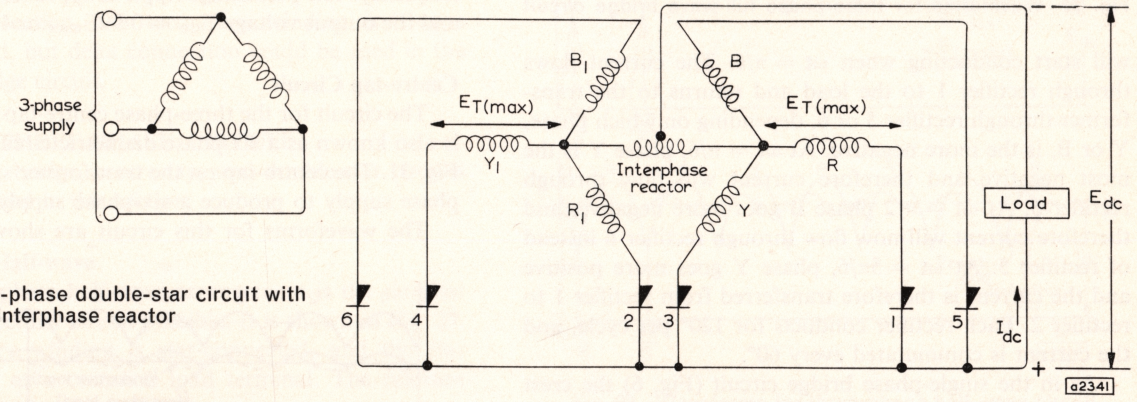


Fig. 33—Three-phase double-star circuit with interphase reactor

At any instant, current is carried by two phases – one in each star – as shown in Fig. 34. The return current is divided between the two secondaries by the interphase reactor. Thus the instantaneous output voltage is the average of the instantaneous voltages of the two secondaries that are conducting. The variation in the d.c. current produces a third harmonic e.m.f. across each half of the interphase reactor, which adds to the e.m.f. of one anode and subtracts from that of the other, thus holding the two at a common voltage. At low d.c. currents, a transition point is reached when the current is too small to produce the third harmonic e.m.f., and the circuit reverts to the three-phase centre-tap system, giving a sudden rise in the output voltage.

The circuit has a six-phase ripple but a three-phase voltage ratio. Its use reduces the line current to approximately half that of the three-phase centre-tap circuit, therefore rectifiers with a smaller peak current rating can be used. However, the peak inverse voltage is somewhat greater.

This arrangement may be used where the cost of the interphase reactor is offset by the use of rectifiers with relatively low current ratings. It is frequently used for low-voltage high-current electrolytic plant. The utility factors of both primary and secondary are high, but that of the secondary is lower by a factor of $\sqrt{2}$ than that of the three-phase bridge circuit.

Double-star Circuit without Interphase Reactor

The purpose of the interphase reactor in the above circuit is to give a third harmonic e.m.f. which allows two rectifiers to conduct at the same time. A similar effect can be produced by using a transformer with a star-connected primary and a centre-tapped star secondary of the type shown in Fig. 31. The two star points must not be connected.

With this arrangement, the transition from three-phase double-star operation to six-phase operation occurs at a higher current, unless special attention is given to the design of the transformer to give a high zero-phase sequence reactance. One method of obtaining this is to use a five-limb core with the windings on the three centre limbs.

The circuit is used for low-voltage electrolytic plant which is unlikely to be operated at currents less than 25% of full load current.

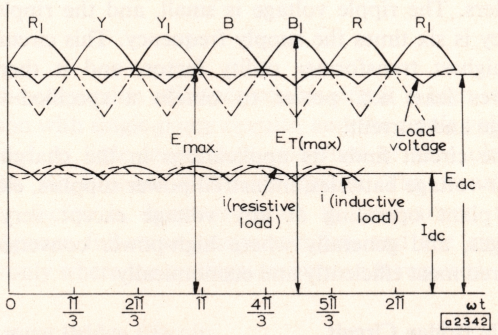


Fig. 34—Waveforms for three-phase double-star circuit with interphase reactor

Smoothing of Three-phase Circuit Output

For the power visualised in three-phase silicon rectifier circuits, it would be prohibitive to use any filter circuit. A choke input filter may be a practical proposition where only a small current is required, but at high currents the size of the shunt capacitor required will be enormous, and it will have to carry a large ripple current.

For the sake of completeness, the values of critical inductance and various other relevant details have been given in Table 5. These values may be derived by a procedure similar to that described on page 248.

IDEALISED ANALYSIS OF POLYPHASE CIRCUITS

The voltage and current relationships, transformer rating, ripple, and interphase reactor rating for polyphase circuits are discussed in this section.

TABLE 5

Choke Input Filter Performance

	Percentage Ripple $V_R\%$		**Critical Inductance** L_{crit}(Henries)		**RMS Ripple Current** $I_{c(rms)}$(amps)	
	General formula	50c/s supply frequency	General formula	50c/s supply frequency	General formula	50c/s supply frequency
Single-phase full-wave	$\frac{1{\cdot}193}{f_r^2LC}$	$\frac{119{\cdot}3}{LC}$	$\frac{R_{LOAD(max)}}{3\pi f_r}$	$\frac{R_{LOAD(max)}}{943}$	$\frac{E_{dc}}{13{\cdot}3f_rL}$	$\frac{E_{dc}}{1330L}$
Three-phase half-wave	$\frac{0{\cdot}45}{f_r^2LC}$	$\frac{20}{LC}$	$\frac{R_{LOAD(max)}}{8\pi f_r}$	$\frac{R_{LOAD(max)}}{3770}$	$\frac{E_{dc}}{35{\cdot}5f_rL}$	$\frac{E_{dc}}{5310L}$
Three-phase full-wave bridge	$\frac{0{\cdot}102}{f_r^2LC}$	$\frac{1{\cdot}133}{LC}$	$\frac{R_{LOAD(max)}}{35\pi f_r}$	$\frac{R_{LOAD(max)}}{33\,000}$	$\frac{E_{dc}}{155f_rL}$	$\frac{E_{dc}}{46\,500L}$

$R_{LOAD(max)}$ in Ω, C in μF, and L in henries

Voltage Relationships

The polyphase rectified output voltage from a sinusoidal supply can be represented by a series:

$$e = E_{max}\frac{N}{\pi}\sin\frac{\pi}{N}\left(1+\frac{2\cos N\theta}{N^2-1}-\frac{2\cos 2N\theta}{4N^2-1}+\frac{2\cos 3N\theta}{9N^2-1}-\ldots\right) \quad \ldots(13)$$

where

E_{max} = peak output voltage

and

N = number of output voltage pulses per cycle of supply voltage.

In the three-phase circuits discussed, N has the following values:

half-wave	$N = 3$
full-wave bridge	$N = 6$
centre-tap	$N = 6$
double-star	$N = 6.$

The following analysis applies to all four of the three-phase circuits that have been discussed. Modifications to the general analysis for particular circuit configurations are stated where necessary.

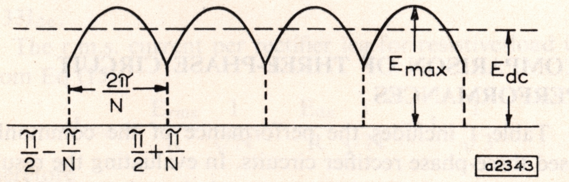


Fig. 35—Output voltage waveform for three-phase circuit

Consider the output voltage waveform shown in Fig. 35, where

$$E_{dc} = \frac{1}{2\pi/N}\int_{\frac{\pi}{2}-\frac{\pi}{N}}^{\frac{\pi}{2}+\frac{\pi}{N}} E_{max}\sin\omega t\,d(\omega t)$$

and therefore

$$E_{dc} = \frac{N}{\pi}E_{max}\sin\frac{\pi}{N}.$$

The value of E_{dc} can also be directly obtained from Eq (13), where the first term represents E_{dc}.

The r.m.s. output voltage is given by

$$E_{rms} = \sqrt{\left(\frac{1}{2\pi/N}E_{max}^2\int_{\frac{\pi}{2}-\frac{\pi}{N}}^{\frac{\pi}{2}+\frac{\pi}{N}}\sin^2\omega t\,d(\omega t)\right)}$$

hence

$$E_{rms} = \frac{\pi}{N}\frac{E_{dc}}{\sin\frac{\pi}{N}}\sqrt{\left[\frac{N}{2\pi}\left(\frac{\pi}{N}+\frac{\sin\frac{2\pi}{N}}{2}\right)\right]} \quad \ldots(14)$$

Current Relationships

For a resistive load,

$$I_{rms}\text{ per rectifier leg} = \frac{E_{rms}}{R}\cdot\frac{1}{\sqrt{N}}.$$

Modifications are necessary for this equation to be valid for all four circuits. If we define I_{rms} as

$$I_{rms} = \frac{E_{rms}}{R}\frac{1}{\sqrt{N'}}, \quad \ldots(15)$$

then from Eq (14)

$$I_{rms} = \left[\frac{\pi}{N}\frac{I_{dc}}{\sin\frac{\pi}{N}}\sqrt{\left(\frac{N}{2\pi}\left(\frac{\pi}{N}+\frac{\sin\frac{2\pi}{N}}{2}\right)\right)}\right]\frac{1}{\sqrt{N'}}. \quad \ldots(16)$$

In this equation, the term $1/\sqrt{N'}$ has a denominator appropriate to the circuit configuration used. Thus, for three-phase half-wave $\sqrt{N'} = \sqrt{3}$, for three-phase full-wave bridge $\sqrt{N'} = \sqrt{3}$, for three-phase centre-tap $\sqrt{N'} = \sqrt{6}$, and for three-phase double-star $\sqrt{N'} = 2\sqrt{3}$.

It should be noted that in the three-phase full-wave bridge circuit a pair of rectifiers conduct at any one instant, therefore $\sqrt{N'} = \sqrt{3}$ and not $\sqrt{6}$.

In the three-phase double-star circuit, the direct current is supplied by two separate star windings, therefore $\sqrt{N'} = 2\sqrt{3}$.

For an inductive load,

$$I_{rms}\text{ per rectifier leg} = I_{dc}/\sqrt{N'} \quad \ldots(17)$$

where the values of $\sqrt{N'}$ given for Eq (16) also apply.

The average current per rectifier leg $I_0 = I_{dc}/N$. This equation is valid for the three-phase rectifier circuits with the exception of the three-phase full-wave bridge. For this circuit, N = 3, because a pair of rectifiers conduct at any one instant.

Transformer Rating

The transformer secondary phase voltage $E_{T(rms)} = E_{max}/\sqrt{2}$. Modifications are necessary for this expression to be valid for all four circuits.

If we define $E_{T(rms)}$ as

$$E_{T(rms)} = \frac{1}{K} \cdot \frac{E_{max}}{\sqrt{2}}$$

then from Eq (13)

$$E_{T(rms)} = \frac{1}{K} \cdot \frac{\pi}{\sqrt{2N}} \cdot \frac{1}{\sin\frac{\pi}{N}} E_{dc} \qquad \ldots(18)$$

where for three-phase half-wave K = 1, for three-phase full-wave bridge $K = \sqrt{3}$, for three-phase centre-tap K = 1, and for three-phase double-star $K = \sqrt{3}/2$.

For the three-phase full-wave bridge circuit, $K = \sqrt{3}$ because the output voltage E_{max} is supplied by the three-phase line voltage.

For the three-phase double-star circuit, $K = \sqrt{3}/2$ because

$$E_{max} = E_{T(max)} \cos 30 = \frac{\sqrt{3}}{2} E_{T(max)}.$$

The transformer secondary r.m.s. current $I_{T(rms)} = I_{rms}$. This expression needs modification for the full-wave rectifier bridge circuit.

If we define $I_{T(rms)}$ as

$$I_{T(rms)} = M . I_{rms} \qquad \ldots(19)$$

then for three-phase half-wave M = 1, for three-phase full-wave bridge $M = \sqrt{2}$, for three-phase centre-tap M = 1, and for three-phase double-star M = 1.

For the three-phase bridge circuit, $M = \sqrt{2}$ because each transformer winding supplies current to the circuit twice per cycle.

The secondary volt-amp rating is

$$VA_s = n(E_{T(rms)} . I_{T(rms)}) \qquad \ldots(20)$$

where n = number of secondary windings.

$$\text{The secondary utility factor} = \frac{E_{dc} . I_{dc}}{VA_s} \qquad \ldots(21)$$

Percentage Ripple

The percentage ripple is given by

$$\% \text{ ripple} = \frac{\text{Fundamental r.m.s. ripple voltage}}{E_{dc}} \times 100.$$

From Eq (13), if ripple frequencies other than the fundamental are ignored,

$$\% \text{ ripple} = \frac{2}{N^2-1} \cdot \frac{1}{\sqrt{2}} 100 = \frac{141}{N^2-1}. \qquad \ldots(22)$$

The fundamental ripple frequency is

$$f_r = Nf, \qquad \ldots(23)$$

where f is the supply frequency.

Rating of Interphase Reactor

Rectification with a double-star circuit requires an interphase reactor. The rating of this reactor can be calculated as follows.

If it is assumed that a triangular waveform appears across the reactor in the process of holding the phase voltages of the two star circuits at a common value, the crest value of this waveform will be V_{max}. The frequency is three times the supply frequency. The voltage across the reactor is at its maximum when the phase voltage of one star connection is at its maximum. The phase voltage of the other star connection is displaced by $\pi/3$, and therefore is at half maximum when the maximum voltage appears across the reactor. Thus

$$V_{max} = E_{T(max)} - \frac{E_{T(max)}}{2}$$
$$= \frac{E_{T(max)}}{2} = \frac{\sqrt{2}}{2} E_{T(rms)}.$$

The triangular waveform can also be represented by a sine series

$$v = V_{max} \frac{8}{\pi^2} \left(\sin\theta - \frac{1}{9} \sin 3\theta + \frac{1}{25} \sin 5\theta - \ldots \right) \qquad \ldots(24)$$

If the third and higher harmonics are ignored, the peak value of an equivalent sinewave is $E_{eq(max)}$. Thus

$$E_{eq(max)} = \frac{8}{\pi^2} V_{max} = \frac{8}{\pi^2} \frac{\sqrt{2}}{2} E_{T(rms)}$$

therefore

$$E_{eq(rms)} = \frac{1}{\sqrt{2}} \left(\frac{8}{\pi^2} \frac{\sqrt{2}}{2} E_{T(rms)} \right) = \frac{4}{\pi^2} E_{T(rms)}.$$

Now the form factor for a triangular waveform is r.m.s./mean = 1·16, therefore the r.m.s. voltage rating of the reactor is

$$\frac{4}{\pi^2} \frac{E_{dc}}{1{\cdot}16}.$$

The current through the reactor is $I_{dc}/2$, therefore the rating of the reactor is

$$\frac{4}{\pi^2 \times 1{\cdot}16 \times 2} E_{dc} . I_{dc} = 0{\cdot}174 E_{dc} . I_{dc}. \qquad \ldots(25)$$

COMPARISON OF THREE-PHASE CIRCUIT PERFORMANCES

Table 1 includes the performance of the commonly-used three-phase rectifier circuits. In evaluating the results in this table, it has been assumed that the transformer and rectifiers are ideal. The table, however, gives a good indication of the relative merits of the circuits, and may be used to select the best circuit for any particular application. It may also be used for comparing the kilowatts per rectifier available from various circuits. This is best illustrated by an example.

Consider the single-phase and three-phase full-wave bridge circuits, and assume that rectifiers rated at a crest working voltage of 400V and with a current rating of 20A are available. The attainable performances are compared in Table 6.

From the above calculation, it follows that better use of rectifiers is made in the three-phase bridge circuit.

TABLE 6

Comparison of Three-phase Circuits

	Single-phase bridge	**Three-phase bridge**
Number of rectifiers in circuit From Table 1	4	6
Output voltage E_{dc}	$\frac{400}{1{\cdot}57} = 255\text{V}$	$\frac{400}{1{\cdot}05} = 380\text{V}$
Output current I_{dc}	$2 \times 20 = 40\text{A}$	$3 \times 20 = 60\text{A}$
Power available $E_{dc} . I_{dc}$	$255 \times 40 = 10{\cdot}2\text{kW}$	$380 \times 60 = 22{\cdot}8\text{kW}$
Kilowatts per rectifier	$\frac{10{\cdot}2}{4} = 2{\cdot}55\text{kW}$	$\frac{22{\cdot}8}{6} = 3{\cdot}8\text{kW}$

Idealised Analysis of Three-phase Bridge Circuit

The three-phase bridge circuit will be analysed as an example of how various values, tabulated in Table 1, are determined.

The output voltage waveform is given by Eq (13). The number N of voltage pulses per cycle of mains voltage, is 6. Therefore

$$e = \frac{3}{\pi} E_{max} \left(1 + \frac{2}{35} \cos 6\theta - \frac{2}{143} \cos 12\theta + \ldots \right)$$

and

$$E_{dc} = \frac{3}{\pi} E_{max} = 0{\cdot}955 E_{max}$$

or

$$E_{max} = \frac{E_{dc}}{0{\cdot}955} = 1{\cdot}05 E_{dc}.$$

From Eq (14)

$$E_{rms} = \frac{\pi}{3} E_{dc} \sqrt{\left(\frac{6}{2\pi} \left(\frac{\pi}{6} + \frac{\sqrt{3}}{4} \right) \right)}$$

therefore

$$E_{rms} = 1{\cdot}0 E_{dc}.$$

The average output current per rectifier leg $= I_{dc}/3 = 0{\cdot}33 I_{dc}$.

The r.m.s. current per rectifier leg for resistive load is, from Eq (15),

$$I_{rms} = \frac{E_{rms}}{R} \cdot \frac{1}{\sqrt{N'}} = \frac{E_{dc}}{R} \frac{1}{\sqrt{N'}} = I_{dc} \frac{1}{\sqrt{3}}$$

therefore

$$I_{rms} = 0{\cdot}577 I_{dc}.$$

With an inductive load

$$I_{rms} = \frac{I_{dc}}{\sqrt{N'}} = 0{\cdot}577 I_{dc}.$$

The peak current per rectifier leg for a resistive load is

$$I_{pk} = \frac{E_{max}}{R} = \frac{1{\cdot}05 E_{dc}}{R} = 1{\cdot}05 I_{dc},$$

and for an inductive load

$$I_{pk} = I_{dc}.$$

The transformer secondary r.m.s. current is, from Eq (19),

$$I_{T(rms)} = \sqrt{2} I_{rms} = 0{\cdot}816 I_{dc}.$$

The transformer secondary r.m.s. voltage is, from Eq (18),

$$E_{T(rms)} = \frac{1}{\sqrt{3}} \cdot \frac{\pi}{6\sqrt{2}} 2E_{dc} = 0{\cdot}428 E_{dc},$$

therefore the total transformer secondary volt-amp rating is, from Eq (20),

$$\begin{aligned} VA_s &= (E_{T(rms)} . I_{T(rms)})3 \\ &= (0{\cdot}428 \times 0{\cdot}816)3E_{dc} . I_{dc} \\ &= 1{\cdot}05 E_{dc} . I_{dc}. \end{aligned}$$

The secondary utility factor, from Eq (21), is

$$U_s = \frac{E_{dc} . I_{dc}}{1{\cdot}05 E_{dc} . I_{dc}} = 0{\cdot}95.$$

Since the line current is symmetrical in this circuit,

Primary r.m.s. phase current = secondary r.m.s. phase current × turns ratio $= 0{\cdot}816 I_{dc}(N_s/N_p)$.

$$\text{The primary r.m.s. phase voltage} = E_{T(rms)} \frac{N_p}{N_s}$$
$$= 0{\cdot}428 E_{dc} \frac{N_p}{N_s}$$

therefore

$$\text{primary volt-amp rating } VA_p = 0{\cdot}816 I_{dc} \frac{N_s}{N_p} \times 0{\cdot}428 E_{dc} \frac{N_p}{N_s}$$
$$= 1{\cdot}05 E_{dc} . I_{dc}.$$

$$\text{Primary utility factor} = \frac{E_{dc} . I_{dc}}{1{\cdot}05 E_{dc} . I_{dc}} = 0{\cdot}95.$$

Fundamental ripple frequency = 6f (from Eq (23)).

$$\text{Percentage ripple } V_R\% = \frac{141}{6^2-1} = 4{\cdot}03\% \text{ (from Eq (22))}.$$

$$\text{Crest working voltage} = 2E_{T(max)} \cos\frac{5\pi}{6}$$
$$= 2\sqrt{2} \frac{\sqrt{3}}{2} E_{T(rms)} = 2{\cdot}45 E_{T(rms)},$$

and

$$\text{crest working voltage} = 2{\cdot}45 \times 0{\cdot}428 E_{dc} = 1{\cdot}05 E_{dc}.$$

LOSSES IN THREE-PHASE CIRCUITS

The output voltage in a practical rectifier circuit is lower than the ideal value because of regulation. The voltage regulation of the three-phase system depends on three factors: the copper loss of the transformer, the rectifier voltage drop, and the commutation voltage drop.

Copper Loss

The reduction in output voltage due to transformer copper loss can be calculated as follows:

Voltage drop due to copper loss E_K=

$$\frac{\text{Transformer copper loss in watts } P_K}{I_{dc}}. \qquad \ldots(26)$$

The value of P_K may be obtained from the short-circuit test on the transformer (Ref. 7).

Rectifier Forward Voltage Drop

The loss due to the forward voltage drop of the rectifier is generally small, especially with silicon rectifiers, which have a voltage drop of only one or two volts. An accurate value for any particular type can be obtained from the forward voltage/forward current characteristic of the rectifier.

The effect of this loss will depend on how many rectifiers are used in series. In particular, it should be noted that in any bridge circuit with one rectifier per leg, the forward voltage drop is that due to two rectifiers in series.

Commutation Loss

The inductance of the transformer winding prevents the current from transferring instantaneously from one phase to the next. Thus for a period the two rectifiers conduct simultaneously. During the commutation period the rectifier output voltage is the average of the instantaneous voltages of the two phases; therefore the output voltage is reduced by the shaded area shown in Fig. 36.

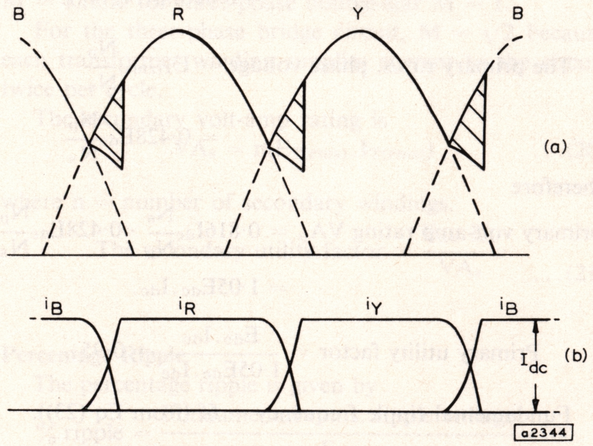


Fig. 36—Commutation loss due to transformer reactance
(a) voltage (b) current

During the commutation period the d.c. current is the sum of the increasing current of the oncoming rectifier and the decreasing current of the previously conducting rectifier. The commutation period ends when the current through the conducting rectifier falls to zero, since it cannot pass any current in the reverse direction.

The voltage drop due to commutation, E_{com}, increases with an increase in the number of phases and an increase in load current. In order to keep the commutation loss to a minimum, the commutating reactance must be limited to a small value, bearing in mind that under short-circuit conditions, the short-circuit current which is limited by the commutating reactance must not exceed the surge current rating of the rectifier. The commutation loss is given by

$$E_{com} = \frac{nX_LI_{dc}}{2\pi}. \qquad \ldots(27)$$

where

n = number of phases
X_L = transformer reactance per phase
I_{dc} = d.c. load current.

With all these losses taken into account, the direct voltage appearing across the load is $E_{dc(actual)}$:

$$E_{dc(actual)} = E_{dc(idealised)} - \frac{nX_LI_{dc}}{2\pi} - \frac{P_K}{I_{dc}} - V_D \times$$
$$\times \text{ number of rectifiers in series}$$

where V_D is the forward voltage drop per rectifier.

EXAMPLES ON THREE-PHASE RECTIFIER CIRCUITS

Consider a three-phase rectifier circuit to supply 90A. Suppose the available rectifiers have a maximum crest working voltage rating of 400V. The transformer has a percentage reactance of 5% and a copper loss of 900W.

The design of suitable bridge and centre-tap circuits is summarised in Table 7.

TRANSFORMER REACTANCE AND CIRCUIT EFFICIENCY

The regulation of the three-phase rectifier circuits depends mainly on the transformer performance. In order to estimate the transformer performance, it is necessary to carry out an open-circuit and short-circuit test on the transformer. The purpose of each is briefly outlined below; but for details of these tests, appropriate literature should be consulted (Ref. 7).

Open-circuit Test

With either the primary or secondary open-circuited, the current and power at normal voltage and frequency are measured. The current $I_{o/c}$ is the sum of the magnetising current and core loss components. The power indicated, $W_{o/c}$, represents the core loss and copper loss. The latter is small and therefore may be neglected, since $I_{o/c}$ is small compared with the full-load current.

Short-circuit Test

In the short-circuit test, either primary or secondary is short-circuited, and the voltage is gradually increased to circulate the rated current through the winding. The short-circuit voltage $V_{s/c}$ necessary to circulate the full load current is measured. The power reading $W_{s/c}$ in this test represents the copper loss I^2R and a small core loss that may be ignored.

TABLE 7

Design of Three-phase Circuits

	Three-phase bridge	Three-phase centre-tap
From Table 1		
Maximum a.c. r.m.s. phase voltage that may be applied	400/2·45 = 163V	400/2·83 = 141·4V
Idealised output direct voltage	163/0·428 = 380V	141·4/0·74 = 191·2V
Average current per rectifier leg	90/3 = 30A	90/6 = 15A
Suitable rectifier	BYY15	BYX13–800
Total rectifier voltage drop at average current	0·98×2 = 1·96V	0·96V
Voltage drop due to commutation overlap (Eq(27))	$\frac{3\times0{\cdot}05\times163}{2\pi} = 3{\cdot}9\text{V}$	$\frac{6\times0{\cdot}05\times141{\cdot}4}{2\pi} = 6{\cdot}75\text{V}$
Voltage drop due to copper loss	900/90 = 10V	900/90 = 10V
approximate voltage available at output terminals	380−1·96−3·9−10 = 364V	191·2−0·96−6·75−10 = 173V
Output power $E_{dc}.I_{dc}$	364×90 = 32·8kW	173×90 = 15·6kW

Calculation

From the above two tests, the performance of the transformer may be calculated as follows:

Transformer rating	M volt-amps
Transformer connection	Delta-star
Normal primary voltage	E_p volts
Normal secondary voltage	E_s volts

Open-circuit test on star side at normal voltage E_s

Core loss = $W_{o/c}$ No-load current = $I_{o/c}$

Short-circuit test on delta side with secondary short-circuited

Short-circuit voltage $V_{s/c}$ Copper loss $P_K = W_{s/c}$ watts at rated current

$$\text{Primary line current} = I_p = \frac{M}{\sqrt{3}E_p}$$

Primary phase current = $I_p/\sqrt{3}$

$$\text{Copper loss per phase} = \frac{W_{s/c}}{3} \text{ watts}$$

Reactance e.m.f. per phase at current $I_p/\sqrt{3} = E_x$

$$= \sqrt{\left(V_{s/c}^2 - \left(\frac{W_{s/c}}{3}\cdot\frac{\sqrt{3}}{I_p}\right)^2\right)}$$

$$\text{Therefore \% reactance} = \frac{E_x}{E_p}100 = X\% \qquad \ldots(29)$$

Circuit Efficiency

$$\%\text{ efficiency} = \frac{\text{Output}}{\text{output}+\text{losses}}\,100$$
$$= \left(1 - \frac{\text{losses}}{\text{output}+\text{losses}}\right)100 \qquad \ldots(30)$$

where total losses = $W_{o/c} + W_{s/c} + I_{dc}.V_D \times$ number of rectifiers in series.

General Notes on Rectifier Use

In this final section, the operation of silicon diode rectifiers in parallel and in series is considered, as well as the general question of the protection of rectifiers against circuit failures and transient voltages.

RECTIFIERS IN PARALLEL

Rectifiers are connected in parallel when the current requirement exceeds the current rating of a single rectifier. For very large currents it may be necessary to parallel more than two rectifiers.

The spreads in the forward characteristic of the rectifiers lead to unequal distribution of current between the rectifiers when connected in parallel. This may cause the current and temperature ratings of the rectifiers to be exceeded.

The following methods or combination of methods may be used to make the rectifiers share the current. These methods will be discussed in detail in a later article (Ref. 8).

Current Derating

The rectifiers are derated so that the rectifier having the best forward characteristic will not exceed the rated current when it is connected in parallel with rectifiers having the worst characteristic.

Temperature Derating

The rectifiers may be overloaded, provided that the thermal resistance from mounting base to ambient is reduced so that the maximum temperature rating is not exceeded.

Thermal Coupling

The divergence in the forward characteristic is minimised by mounting the rectifiers on the same heatsink, so that the mounting base temperatures of the rectifiers will be equalised.

Series Resistors

The rectifiers can be made to current-share by inserting a small resistance in series with each rectifier. This will tend to reduce the spread in the forward characteristic.

Balancing Transformers

Coils which have a common core may be inserted in series with the rectifiers, so that the induced e.m.f. produced by the difference in currents will cause the currents in the rectifiers to balance.

Selection of Rectifiers

The rectifier currents are balanced by limiting the spread of the forward characteristic.

RECTIFIERS IN SERIES

Rectifiers are connected in series when the output voltage requirement exceeds the voltage rating of a single rectifier. For high-voltage applications it may be necessary to connect two or more rectifiers in series.

In the forward direction, each rectifier carries a common current. In the blocking state, the voltage across each rectifier will depend on the leakage current. Since there is a spread in the leakage current rating of the rectifiers, it is necessary to connect a resistor R across each rectifier to ensure that the rated crest working voltage is not exceeded.

If the rectifier is reverse-biased immediately after it has been carrying forward current, it requires a finite time to recover to its blocking state. This time will depend on the stored charge, which varies from one rectifier to another. Thus, if a reverse voltage is applied across several rectifiers in series, then the rectifier with the least stored charge will recover first and will be subjected to the full applied voltage. This may damage one or more rectifiers. A capacitor C is connected across each rectifier for protection against transient voltages.

Where only two rectifiers are in series, the capacitors are not always needed, because the instantaneous value of applied voltage, immediately after blocking, frequently does not exceed the voltage rating of one rectifier.

Determination of Reverse Voltage Sharing Resistor

A chain of n rectifiers connected in series is shown in Fig. 37a. Across each is connected a resistor R with a tolerance ±ΔR.

Consider rectifier 1 to be an ideal rectifier (that is, with no leakage current) and the rest of the rectifiers to have maximum leakage current $I_{R(max)}$ at maximum junction temperature. The maximum voltage V_1 to which rectifier 1 will be subjected will occur when the resistor across rectifier 1 is $(R+\Delta R)$ and that across each of the other rectifiers is $(R-\Delta R)$. This condition is shown in Fig. 37b, where the voltage across rectifier 1 is

$$V_1 = (I_1+I_{R(max)})(R+\Delta R) \qquad \ldots(31)$$

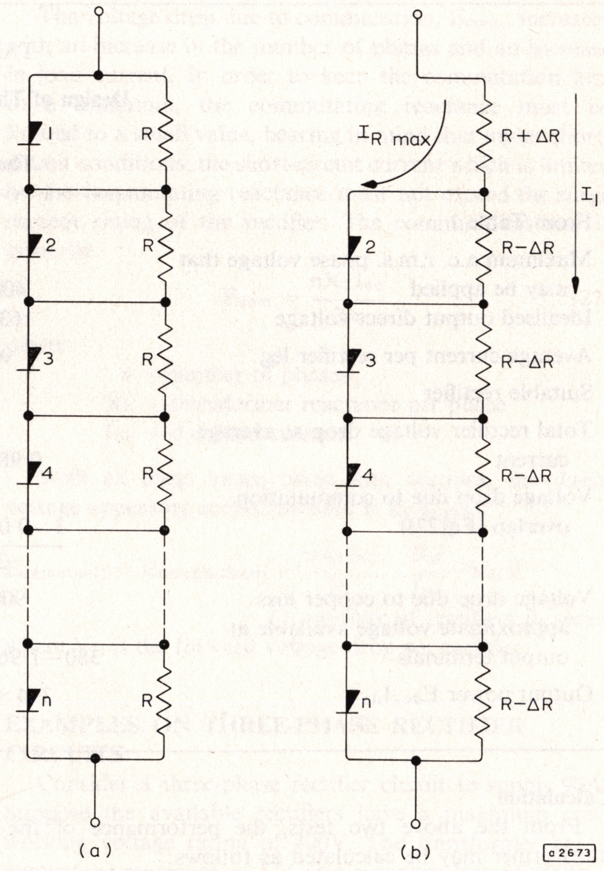


Fig. 37—Determination of voltage-sharing resistor

and the voltage across the rest of the chain is

$$V_2 = I_1(R-\Delta R)(n-1) \qquad \ldots(32)$$

therefore

$$I_1 = \frac{V_2}{(R-\Delta R)(n-1)}. \qquad \ldots(33)$$

Substitution of (33) in (31) gives

$$V_1 = \left[\frac{V_2}{(R-\Delta R)(n-1)} + I_{R(max)}\right](R+\Delta R) \qquad \ldots(34)$$

and multiplying (34) by $\frac{R}{R+\Delta R}$ gives

$$\frac{V_1 R}{R+\Delta R} = \frac{V_2 R}{(R-\Delta R)(n-1)} + I_{R(max)}R. \qquad \ldots(35)$$

If $\beta = \Delta R/R$, Eq (35) then becomes

$$\frac{V_1}{1+\beta} = \frac{V_2}{(1-\beta)(n-1)} + I_{R(max)}R. \qquad \ldots(36)$$

Since V_1+V_2 = applied reverse voltage V_R

$$R = \frac{1}{I_{R(max)}}\left[\frac{V_1}{1+\beta} - \frac{V_R-V_1}{(1-\beta)(n-1)}\right]. \qquad \ldots(37)$$

The maximum value of V_1 that is permissible is the crest working voltage V_{RW} rating of the rectifier, therefore

$$R \leqslant \frac{1}{I_{R(max)}}\left[\frac{V_{RW}}{1+\beta} - \frac{V_R-V_{RW}}{(1-\beta)(n-1)}\right]. \qquad \ldots(38)$$

For this equation to yield practical values of R,

$$\frac{V_{RW}}{1+\beta} > \frac{V_R-V_{RW}}{(1-\beta)(n-1)}$$

or

$$n-1 > \frac{V_R-V_{RW}}{V_{RW}} \cdot \frac{1+\beta}{1-\beta}$$

therefore

$$n > 1 + \frac{V_R-V_{RW}}{V_{RW}} \cdot \frac{1+\beta}{1-\beta}. \qquad \ldots(39)$$

Eq (39) indicates the minimum possible number of rectifiers that has to be connected in series for a given resistor tolerance β.

Example

It is wished to determine the number of BYZ10 rectifiers required in series, and the value of R to be connected across each rectifier, for three-phase full-wave operation in which each arm of the bridge is subjected to a peak reverse voltage V_R of 3600V.

For BYZ10 rectifiers, use V_{RW} = 400V and $I_{R(max)}$ = 600μA. Let the tolerance on the resistor R be ±5%.

From Eq (39)

$$n > 1 + \frac{3600-400}{400} \cdot \frac{1+0{\cdot}05}{1-0{\cdot}05} > 9{\cdot}88.$$

Let n = 10. Then, from Eq (38),

$$R \leqslant \frac{1}{600\times10^{-6}}\left[\frac{400}{1+0{\cdot}05} - \frac{3600-400}{(1-0{\cdot}05)9}\right] \leqslant 10\text{k}\Omega.$$

The power dissipated, P_R, in the resistor R in a three-phase full-wave bridge is given by

$$P_R \simeq \frac{0{\cdot}402 V_{RW}^2}{R}$$

$$\simeq \frac{0{\cdot}402(400)^2}{10\times10^3} \simeq 6{\cdot}43\text{W}.$$

If this dissipation is not tolerable, n must be increased. Let n = 11, then

$$R \leqslant \frac{1}{600\times10^{-6}}\left[\frac{400}{1+0{\cdot}05} - \frac{3600-400}{(1-0{\cdot}05)10}\right] \leqslant 73{\cdot}3\text{k}\Omega.$$

Choose R = 68kΩ±5%. The power dissipated in R is then

$$P_R \simeq \frac{0{\cdot}402(400)^2}{71{\cdot}4\times10^3} \simeq 0{\cdot}9\text{W}.$$

Thus, series connection of eleven BYZ10 rectifiers in each arm of the bridge, with 68kΩ±5% across each rectifier, gives a suitable solution.

Determination of Capacitor for Transient Voltage Sharing

Fig. 38a shows rectifiers connected in a series chain. Across each is connected a capacitor C to protect the rectifiers against voltage transients.

The worst condition will be when rectifier 1 has minimum stored charge Q_{min}, and the rest have maximum stored charge Q_{max}. When a reverse voltage is applied to the chain immediately after forward conduction, the rectifier with the least stored charge will recover first and will be subjected to some voltage V_1. This voltage V_1 must not exceed the maximum repetitive peak inverse

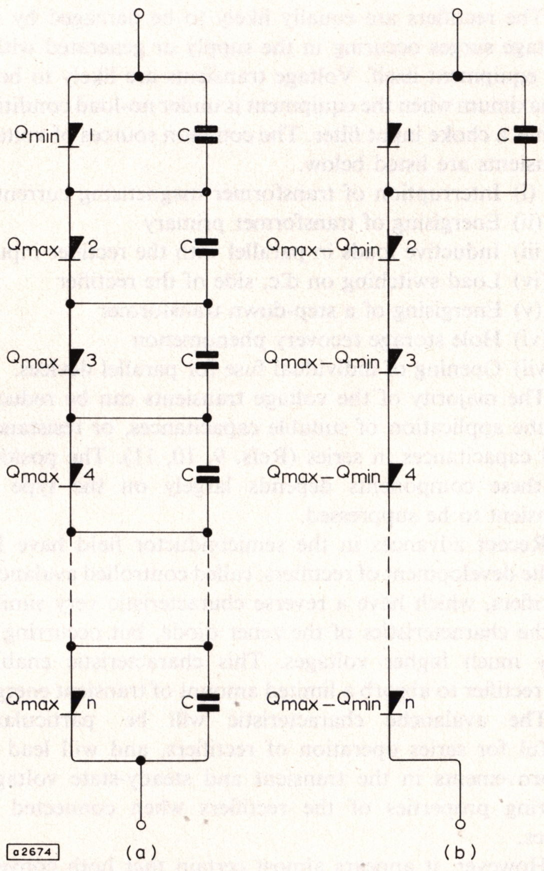


Fig. 38—Determination of transient voltage-sharing capacitor

voltage rating V_{RR} of the rectifier. This condition is shown in Fig. 38b.

Now,

$$Q_{max}-Q_{min} = C\,V_1$$

and since $V_1 \leqslant V_{RR}$

$$C \geqslant \frac{Q_{max}-Q_{min}}{V_{RR}}. \qquad \ldots(40)$$

In deriving the above expression it has been assumed that the recombination time of the rectifiers is about the same.

When the rest of the rectifiers have recovered, the transient voltage will be shared equally, provided that the value of C determined in Eq (40) is at least ten times the junction capacitance. Usually, the value of C given by Eq(40) is very large compared with the junction capacitance.

RECTIFIER PROTECTION

Protection of the rectifier circuit must be carefully designed, as in certain circuits failure of one device may lead to the destruction of the rest of the rectifiers in the circuit (Ref. 9).

To prevent damage to rectifiers in the event of short-circuit, they must be protected by fuses. The fuse must blow before the surge current rating of the rectifier is exceeded; therefore the correct fuse must be selected, due attention being paid to the surge current rating of the rectifier (Refs. 9, 10).

The rectifiers are equally likely to be damaged by the voltage surges occuring in the supply or generated within the equipment itself. Voltage transients are likely to be at a maximum when the equipment is under no-load condition or has a choke input filter. The common sources of voltage transients are listed below.

(i) Interruption of transformer magnetising current
(ii) Energising of transformer primary
(iii) Inductive loads in parallel with the rectifier input
(iv) Load switching on d.c. side of the rectifier
(v) Energising of a step-down transformer
(vi) Hole storage recovery phenomenon
(vii) Opening of individual fuse for parallel devices.

The majority of the voltage transients can be reduced by the application of suitable capacitances, or resistances and capacitances in series (Refs. 9, 10, 11). The position of these components depends largely on the type of transient to be suppressed.

Recent advances in the semiconductor field have led to the development of rectifiers, called controlled avalanche rectifiers, which have a reverse characteristic very similar to the characteristics of the zener diode, but occurring at very much higher voltages. This characteristic enables the rectifier to absorb a limited amount of transient energy.

The avalanche characteristic will be particularly useful for series operation of rectifiers, and will lead to improvements in the transient and steady-state voltage-sharing properties of the rectifiers when connected in series.

However, it appears almost certain that both conventional diode rectifiers and the controlled avalanche types will be needed, since the power absorption capability of the controlled avalanche rectifier is limited, and is not adequate to absorb all the energy contained in voltage surges occurring on normal low-impedance mains supplies.

SUMMARY

In this article an attempt has been made to illustrate the great potentiality of silicon power rectifiers.

For single-phase rectification, the choke input filter is more suitable than the simple capacitor filter, especially where the voltage regulation is of importance. However, up to a certain power level it is feasible to use the simple capacitor filter, particularly where a d.c. supply of one specified current is required, and where the weight of the equipment has to be kept to a minimum. The limiting factor is the maximum ripple current that the capacitor can handle.

Three-phase rectifier circuits can be effectively used for delivering large amounts of power, provided that adequate protection of the rectifiers is included in the circuit. The increased reliability, low running costs, and relatively low cost of the silicon rectifiers, have made it an economical proposition to consider them for heavy current applications.

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